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Current Overload Protection in Thyristor Phase Control Systems

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Thyristors may be used to control the power supplied to either an a.c. or d.c. load from the a.c. mains. With certain types of load, in particular with d.c. shunt motors, the load impedance may be very low under certain operating conditions, such as when starting. To avoid excessive currents some circuit arrangement is desirable to limit this current to an acceptable level. The nature of the problem and some practical circuits are described.

INTRODUCTION

The method of power control from an a.c. supply known as phase control makes use of the principle of allowing current to flow for only a fraction of each half cycle of the supply. The smaller the fraction of a cycle during which current flows, the smaller the power supplied to the load. Thyristors may be used to perform the role of synchronous switch; the power reaching the load is then controlled by selecting the correct angle during each cycle at which to fire the thyristors.

The simplest systems are designed so that the power control thyristors are capable of handling the maximum current which is ever likely to flow under the conditions of minimum load impedance. Where the load impedance may be very low under certain transient conditions, this design principle may result in the need for a very large thyristor stack in order to handle these large transient currents, whereas only a relatively small stack is required for normal running conditions.

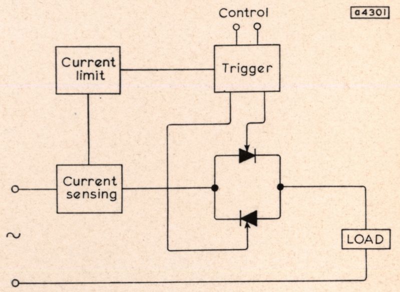
As thyristors are used to control power to the load during normal operation they may also be used to limit current to a predetermined maximum value under abnormal or starting conditions. To do this the load current may be sampled and this information then used in a feedback circuit which controls the firing angle of the thyristors and keeps the current to the selected maximum value.

In addition to providing protection for the power control devices, so that these do not have to be chosen to handle the maximum possible current, the load itself is protected from possible damage.

PHASE CONTROL SYSTEMS

Only four types of phase control system will be considered here: single-phase a.c. control, 3-phase a.c. control, single-phase d.c. control, and 3-phase d.c. control. The basic circuits are shown in Figs. 1, 2, 3, and 4.

Fig. 1—
Single-phase
a.c. control



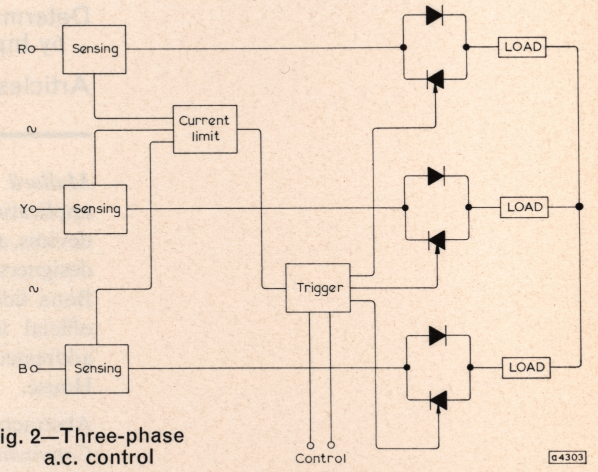


Fig. 2—Three-phase
a.c. control

Any other phase control circuit will be very similar to one of these four types, and it should be a simple matter to adapt one of the circuits to be described to suit any other system. With all these circuits the power to the load may be varied by applying a variable voltage to, or varying the resistance between, the control terminals of the trigger module which supplies trigger pulses to the thyristors. For the circuit to be completely safe under all conditions of operation it must be possible to connect the a.c. supply, with the load connected and the trigger circuit set to the condition of maximum current, without the resulting current exceeding the maximum current rating of the thyristor stack and of the load itself.

To ensure that the maximum safe current is not exceeded the a.c. line current can be measured by using a current transformer or some similar technique. The output voltage from the current transformer is compared with some reference, and when it exceeds this reference voltage the normal control of phase angle by the trigger circuit is overridden, so that the thyristor firing angle is retarded to limit the current to the load.

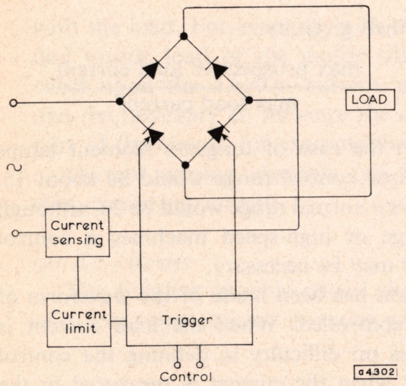


Fig. 3—
Single-phase
d.c. control

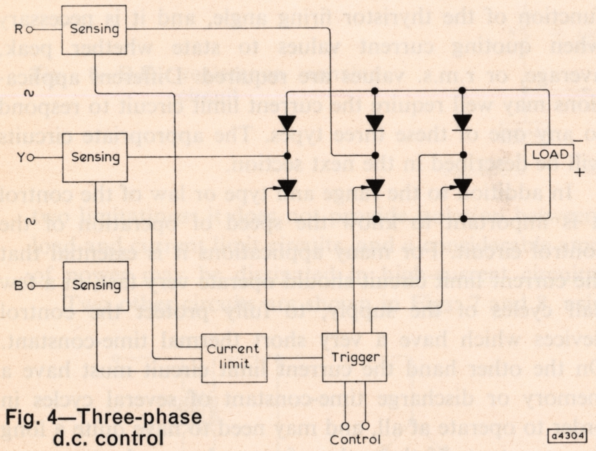


Fig. 4—Three-phase
d.c. control

In the case of the d.c. circuits, Figs. 3 and 4, it may be more convenient to measure the current in the load circuit by means of a low-value series resistance, but in general this will only be applicable to relatively low-power circuits. Alternatively, a saturable reactor or Hall effect device can be used to measure the d.c. current and at the same time provide isolation.

Whichever power control system is used, the basic current limit circuit remains the same. The input signal may be derived from a single- or three-phase a.c. supply, or from a d.c. supply; but when it has been converted to a d.c. voltage for comparison with a reference voltage, the remainder of the circuit has to accept a d.c. difference signal which is then used to control firing angle phase delay.

TYPES OF LOAD AND THEIR CHARACTERISTICS

Tungsten Lamps

A limited number of measurements on 240V tungsten filament lamps between 100W and 1000W rating show that the ratio of hot to cold resistance lies between 13:1 and 15:1. It follows, therefore, that the initial current on switching on a lamp circuit may be very high. Fortunately, the heating time-constant of lamps is small. The filament of a low-wattage lamp will reach working temperature in about one cycle of the supply, so that the large starting surge produces only a transient temperature rise in the control devices. These starting transients may, however, have a large enough energy content to require the use of larger power devices than would be necessary to handle the normal running load current. It may in many instances be desirable to suppress the current surge to effect an overall economy in design, and in any case the reduction of this surge is likely to prolong the life of the lamps.

Motors

The remarks that follow apply primarily to d.c. shunt-wound machines, as this is the type to which speed control is most commonly applied; but to some extent the same will apply to series motors where the winding resistance is particularly low, such as high-starting-torque machines. The starting current of commutator motors is many times the normal running current, particularly with large or high-speed machines. With normal d.c. drives some starting arrangements have to be provided to limit this initial current surge to prevent damage to the motor while it is accelerating. When the motor is being fed through a thyristor control circuit the large starting current would also damage the thyristors and rectifiers, unless they were chosen to have a current rating considerably in excess of that required for normal running. This would result in an unnecessarily bulky and expensive stack. Some protection also has to be provided to safeguard the control devices and motor windings in the event of the motor being stalled.

Typical starting currents are about ten times normal full-load running current for motors of around 1 h.p. or for high-speed special f.h.p. machines. Starting current may be even higher for very-high-speed motors or for larger machines. This current may remain at a high level for several seconds while the motor is accelerating, especially if the mechanical load has appreciable inertia. In such cases some form of current limiting is an essential part of the control system.

Ovens and Furnaces

A number of different materials are used for the heating elements of electric furnaces, depending upon the type of furnace and the temperature range to be covered. Some of these materials, such as platinum, molybdenum, tungsten, and tantalum, have a large temperature coefficient, so that the current taken from the supply when the furnace is cold may be several times the normal running current. Other materials, such as silicon carbide, have a minimum resistance at some intermediate temperature. Materials with a large temperature coefficient may require some current-limiting circuit to be included in the power control circuit so that the current taken during the furnace heating period shall not be excessive. As a guide when designing current limit systems, typical characteristics of certain materials are given in Fig. 5 (p. 264). Minimum element resistance is given as unity, with resistance at other temperatures as a multiple of this value. It should be noted that nickel-chrome has a nearly constant resistance over its usable range.

Inverters

Inverters may be supplied from the 50c/s a.c. mains through a thyristor control bridge where a variable voltage output is required from the inverter, for example when the inverter is used to supply a variable-frequency drive

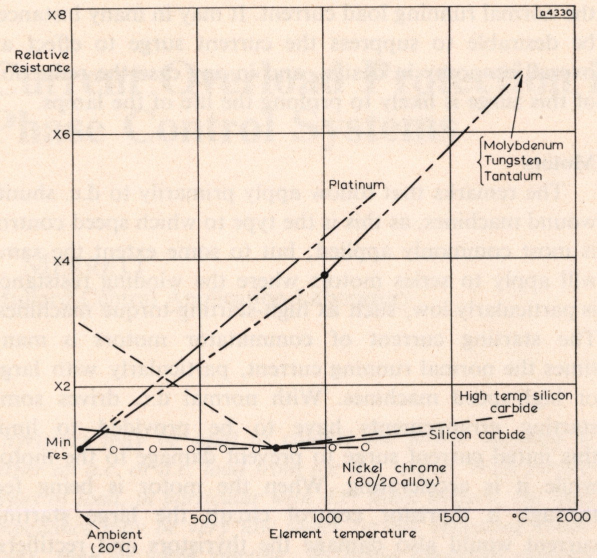


Fig. 5—Relative resistances of furnace element materials throughout their working temperature ranges

NOTE: relative resistances are shown at ambient (switch-on) temperature, at the minimum-resistance point (for those materials that do not have minimum resistance at ambient), and at the maximum working temperature of the material. Relative resistances at other temperatures cannot necessarily be read off; nor can comparisons be made of the actual resistances of the different materials.

for an induction motor. In such cases a current-limit circuit may be required, and is used to control the output of the 50c/s rectifier, as in this way the power input to the inverter is reduced as the inverter load impedance falls.

The design of the current-limit circuit can be similar to that for other types of load, but the load current sensing element can be inserted in the output circuit of the inverter if required. There may be some advantage in doing this if the inverter frequency is higher than 50c/s, as a current transformer can be used with a shorter time-constant smoothing circuit following the normal rectifier to get faster response to overloads.

Any short-circuit protection which may be provided should always be fed from a sensing element in the a.c. mains input, or in the d.c. input to the inverter, if it is to function in the event of an inverter thyristor failing to turn off.

CURRENT-LIMIT SYSTEMS

Current-limit systems are understood here to mean control systems which prevent the load current from rising above some predetermined level when the load impedance falls and the current would normally rise. The current is prevented from rising by delaying the firing of the control thyristors so that the effective supply voltage is reduced as the load impedance falls. In practice it is difficult to design circuits to work satisfactorily over very wide voltage ranges, so it is convenient to specify definite design limits.

These limits can be described in terms of the maximum output current (that is, when the current limit action starts) and the theoretical or 'prospective currents' which would flow in the absence of the current limit circuit when the load has its lowest normal impedance. The circuit operating range is then given by:

$$\text{Control range} = \frac{\text{max prospective load current}}{\text{max load current}}.$$

For example, in the case of tungsten filament lamps the minimum required control range would be about 15. In the case of motors a normal range would be 20, although in the case of large or high-speed machines a control range of at least 40 may be necessary.

So far no mention has been made of the waveform of the current being controlled. When the load current is smooth d.c. there is no difficulty in defining the control characteristics; but when the current is measured in the a.c. line of a phase control system the waveform is a function of the thyristor firing angle, and it is necessary when quoting current values to state whether peak, average, or r.m.s. values are required. Different applications may well require the current limit circuit to respond to any one of these three types. The appropriate circuits will be described in the next section.

In addition to the range and type or law of the control it is important to know the speed of operation of the control circuit. For many applications it is essential that the current limit circuit should operate very fast, in a few half cycles of the supply, to fully protect the control devices which have a very short thermal time-constant. On the other hand the current limit circuit must have a memory or discharge time-constant of several cycles in order to operate at all, and may need to have quite a long time-constant, of the order of several seconds, to ensure feedback loop stability.

A current limit circuit constitutes a negative feedback loop of which the transfer coefficient of the load and thyristor control circuit form a part. It is therefore impossible to design a complete system and ensure stability unless the characteristics of all parts of the system, including the load, are known. One simple principle which can be adopted to allow a circuit to be designed in the case where the complete system is incompletely specified, is to keep the loop gain as low as possible and to provide a very large lumped time-constant which is likely to be several times greater than any other time-constant in the loop. This design principle has two main limitations: a large time-constant would result in a very slow-operating system unless special measures are adopted, and a low gain will give poor control characteristics. However, provided that a moderate control range can be tolerated it is possible to design a current limit circuit which will give a satisfactory performance in a large number of applications.

Current Sensing Elements

In general, loads fall into two groups: those supplied with a rectified phase controlled current and those supplied with phase controlled a.c. In the case of a.c. loads the current is measured in the a.c. line; while for d.c. loads either the a.c. input current or the d.c. load current can be measured. When the load is inductive and a flywheel diode is used across the load it should be noted that the load current may be appreciably greater than the a.c. input current, and the ratio of these currents may vary with the load. For example, with a motor load the speed and torque load of the motor will have a considerable effect upon the a.c./d.c. current ratio. In such cases it may be necessary to measure the current in a particular part of the circuit, for example the armature circuit.

Current sensing elements fall into two groups: d.c. circuits and a.c. circuits.

(a) d.c. circuits

The simplest possible circuit shown in Fig. 6 consists of a low-value resistor in series with the load. This has two limitations: it does not provide isolation between the load and current limit circuits, and a considerable amount of power may be dissipated in high-current circuits.

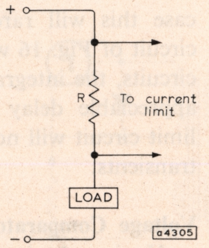


Fig. 6—D.C. resistance sensing circuit

Two other circuits are shown in Figs. 7 and 8, using a saturable reactor and a Hall device. Both of these provide isolation but are relatively expensive, and the component parts are not readily available.

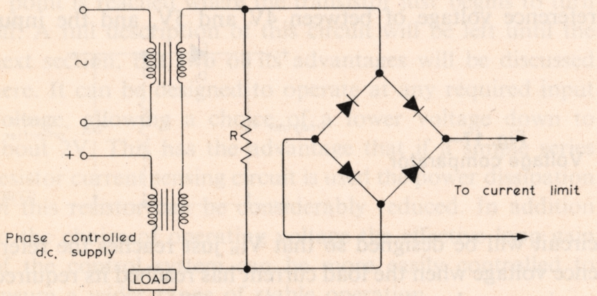


Fig. 7—D.C. saturable reactor sensing circuit

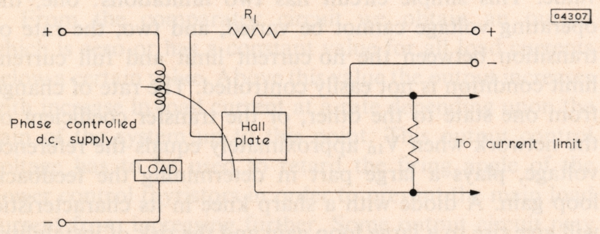


Fig. 8—Hall effect sensing circuit

(b) a.c. circuits

A current transformer and bridge rectifier are normally used in a.c. circuits as shown in Figs. 9 and 10. In the three phase circuit one current transformer can be dispensed with by using the circuit of Fig. 11.

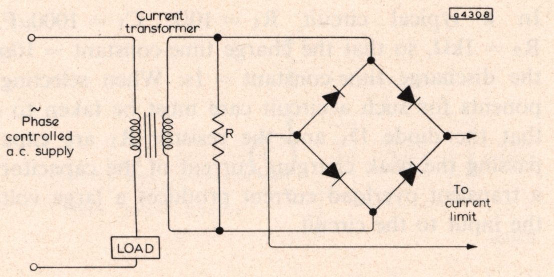


Fig. 9—Single-phase a.c. current transformer sensing

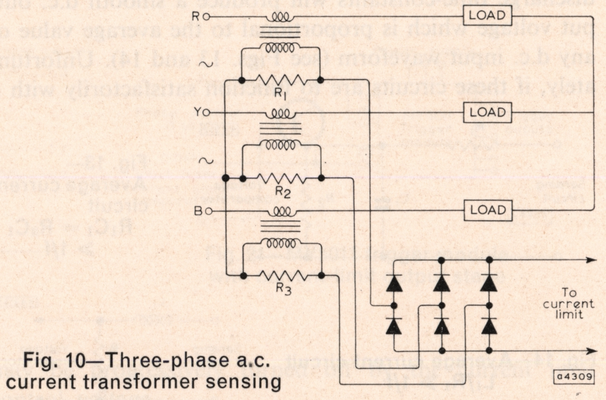


Fig. 10—Three-phase a.c. current transformer sensing

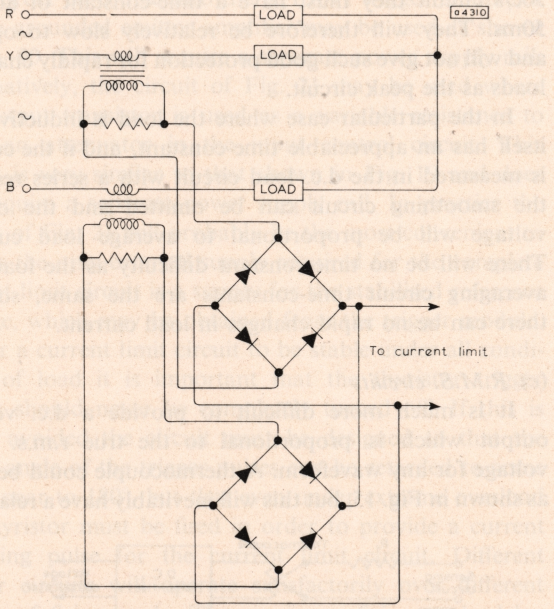


Fig. 11—Three-phase a.c. current transformer sensing

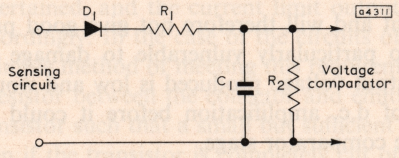


Fig. 12—Peak current circuit $R_2C_1 \gg R_1C_1 \simeq 1/f$

Peak, Average and R.M.S. Circuits

(a) Peak Circuit

The circuit of Fig. 12 with $R_2 \gg R_1$ and $C_1 . R_2 \gg 10$ms (for 50c/s circuits) will provide an output voltage, into a high impedance load, which is approximately equal to the peak input voltage. When used with the current sensing circuits shown in Figs. 7, 9, 10 and 11 the diode D_1 is not required. The response time to changes in load current can be made fast by keeping the value of R_1 low.

In a typical circuit $R_1 = 10\Omega$, $C_1 = 1000\mu F$, and $R_2 = 1k\Omega$, so that the charge time-constant = 10ms and the discharge time-constant = 1s. When selecting components for such a circuit care must be taken to ensure that the diode D_1 and the resistor R_1 are capable of passing the peak charging current of the capacitor when a transient overload current produces a large voltage at the input to the circuit.

(*b*) *Average circuits*

A CR or LR integrating circuit with equal charge and discharge time-constants will produce a smooth d.c. output voltage which is proportional to the average value of any d.c. input waveform (see Figs. 13 and 14). Unfortunately, if these circuits are to function satisfactorily with a 50c/s circuit they must have a time-constant of at least 50ms. They will therefore be relatively slow to operate and will not give such good protection for rapidly changing loads as the peak circuit.

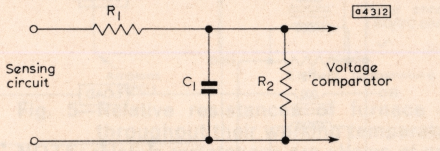


Fig. 13—Average current circuit
$R_1C_1 = R_2C_1 \gg 1/f$

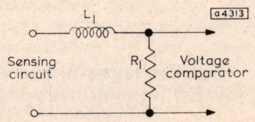


Fig. 14—Average current circuit $L_1/R_1 \gg 1/f$

In the particular case where the load is inductive and itself has an appreciable time-constant, and if the current is measured in the d.c. load circuit with a series resistor, the smoothing circuit can be omitted and the output voltage will be proportional to average load current. There will be no time-constant difficulty as the load and averaging circuit time-constants are the same, so that there can be no rapid changes in load current.

(*c*) *R.M.S. circuits*

It is much more difficult to provide a d.c. voltage output which is proportional to the true r.m.s. input voltage for any waveform. A thermocouple could be used as shown in Fig. 15, but this will inevitably have a relatively long time-constant and will therefore not give good protection. It is also particularly vulnerable to damage by overload. The output voltage produced is low and would require a stage of d.c. amplification before it could be fed to the voltage comparator stage.

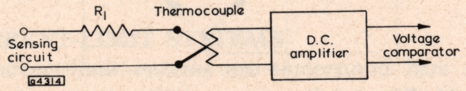


Fig. 15—R.M.S. current circuit

A better but less accurate circuit makes use of the fact that a forward biased diode has an approximately square law over a useful range of current such that $V \propto I^2$. A circuit using this principle is shown in Fig. 16. Having produced a voltage waveform which is proportional to the square of the load current waveform, this has to be integrated to give a d.c. voltage proportional to the mean squared value. If a linear output is required, the square root of this value must be obtained; but in the present case this will rarely be necessary, so that the simple circuit of Fig. 16 will suffice. As with the average current circuits, the integrating circuits unavoidably introduce an appreciable delay into the system so that the current limit circuit will not give protection against rapid current transients.

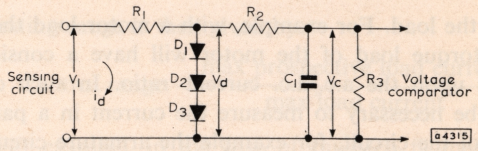


Fig. 16—R.M.S. current circuit

$$R_2C_1 = R_3C_1 \quad i_d \propto V_1 \quad V_d \propto i_d^2$$

$$V_c = \frac{1}{t}\int V_d \, . \, dt = \frac{1}{t}\int V_1^2 dt$$

Voltage Comparators

Having obtained a smooth d.c. voltage, which is proportional to the peak, average, or r.m.s. load current as required, this must now be compared with some reference voltage to determine whether it is above or below the required maximum level. One of the simplest circuits is shown in Fig. 17. The reference diode D_1 will have a reference voltage of between 4V and 5V, and the input circuit will be designed so that V_{in} just reaches the reference voltage when the load current has reached its required maximum value. An output voltage will appear across R_1 when the load current exceeds the designed maximum value. This simple circuit has two limitations: one, the operating voltage cannot be varied, and two, the rate of transition between the no current limit and full current limit condition is not easily controlled. The rate of change from one state to the other, or the transfer coefficient of the network when V_{in} approximately equals the reference voltage, plays a large part in determining the feedback loop gain. A diode with a sharp knee in its characteristic will give rise to a large loop gain and the risk of instability.

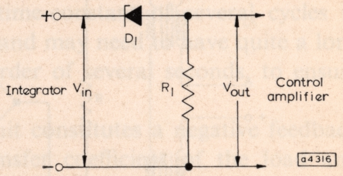


Fig. 17—Voltage comparator

An alternative circuit which effectively uses a low-gain grounded emitter amplifier, with a potentiometer bias supply derived from a stabilised supply as reference, is shown in Fig. 18. With this circuit the input from the current sensing circuit is used to back off the forward bias provided by the base potential divider R_1, R_2, until a point is reached where the transistor just begins to turn off. A full description of this circuit will be left until the next section, but two of its advantages will be discussed here. It can be designed to operate at any required input voltage, allowing a choice of a lower voltage down to about 2V. This has the advantage that if a simple series resistor current sensing circuit is used the power dissipation of this resistor can be considerably reduced. In addition to the choice of operating voltage the effective loop gain of the control system can be more easily controlled to ensure a wider range of stable operation.

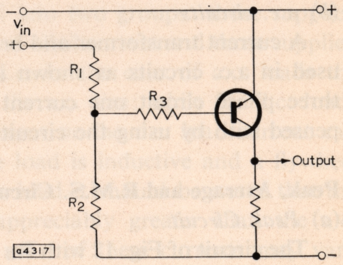


Fig. 18—Voltage comparator

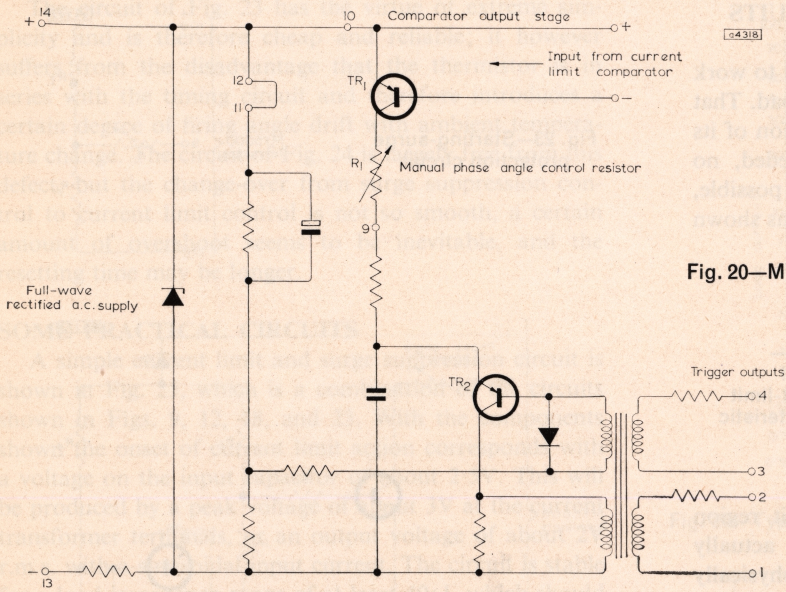

+ 14
10
Comparator output stage
+
Input from current limit comparator
−
12
11
TR1
R1
Manual phase angle control resistor
9
Full-wave rectified a.c. supply
TR2
Trigger outputs
4
3
2
1
− 13
a4318

Fig. 19—MY5000 trigger module with current-limit output stage

Output Circuits

From the comparator circuit we obtain an output which is zero or has a constant value for all load currents below a certain value. Above this value the output increases with increase in load current at a rate depending upon the gain of the system up to this point. This output control voltage has to be used to retard the firing angle of the power control thyristors in order to prevent the main load current from increasing further. Some output circuit may be required to couple the output of the comparator circuit into the input of the thyristor trigger circuit. The nature of this circuit will of course depend upon the trigger circuit being used. Two types of trigger circuit will be considered in detail. A simple blocking oscillator with an RC timing circuit (the Mullard MY5000 trigger module will be used as an example), and a more sophisticated system where the firing angle is controlled by a variable voltage input (for example the Mullard MY5011trigger module).

Considering the simpler circuit first, the basic circuit diagram of the MY5000 is shown in Fig. 19 together with the current limit output circuit. This consists of a transistor TR_1 in series with the variable charging resistor R_1. Transistor TR_1 is normally held in the bottomed state, but may be progressively turned off by the current limit control voltage.

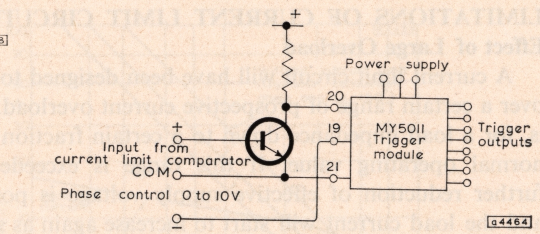


Fig. 20—MY5011 trigger module with current-limit output stage

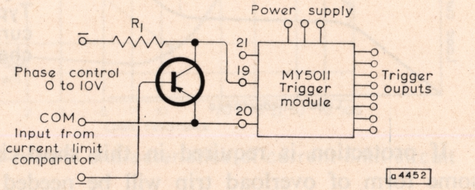


Fig. 21—MY5011 trigger module with current-limit output stage

In order to couple into the MY5011 trigger module, a simple grounded emitter amplifier may be used as shown in Fig. 20. This will require a separate power supply rail. Alternatively, the circuit of Fig. 21 may be used where the output transistor is used as a variable resistance to form an attenuator in the input control circuit. This circuit is very similar to that shown in Fig. 19 and does not require a separate power supply rail. In practice the circuit of Fig. 21 will probably be the most convenient to use, as the only external supply required for the complete current limit circuit is a supply for the stabilised reference voltage, which may be obtained from the trigger module.

For a current limit circuit to be stable under all conditions of load it is important that the thyristor trigger circuit should never be completely turned off. If this is to be avoided the circuit must be designed so that the phase angle of the trigger pulses can be retarded to some angle near to 180°, but that even under extreme overload the thyristor must be fired in order to provide a current sampling pulse for the current limit circuit. Different trigger circuits will operate satisfactorily over different ranges of firing angle. The maximum firing angle delay for the particular trigger circuit being used must be ascertained, and the current limit output circuit designed accordingly. In the case of the circuits of Figs. 19 and 21 a simple method of meeting this requirement is to connect a resistor between the collector and emitter of the output transistor such that a small but sufficient current will flow even if the transistor is completely cut off by a large load current pulse. An alternative arrangement is to connect a resistor between the collector and base of the output transistor. This will serve much the same purpose and at the same time provide some negative feedback to stabilise the gain of the transistor and ensure a wider range of stable operation. A combination of both circuits could be used if necessary.

LIMITATIONS OF CURRENT LIMIT CIRCUITS

Effect of Large Overload

A current limit circuit will have been designed to work over a certain range of prospective current overload. That is, for a load impedance down to a certain fraction of its normal operating value. If this range is exceeded, no further reduction of effective supply voltage is possible, and the load current will start to increase again as shown in Fig. 22.

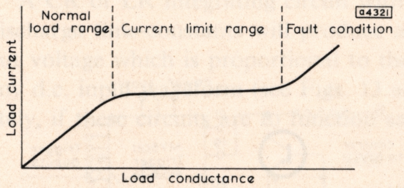


Fig. 22—Typical current-limit characteristic

If protection is required in this short-circuit region some form of overload trip will be needed to actually prevent the thyristor from being fired, or to physically break the circuit. Such a trip will normally remain broken until manually reset.

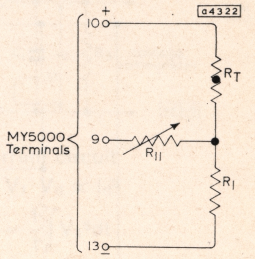
The range of current limit protection will normally only be exceeded under fault conditions, so that some form of resettable trip or fuse will be required to cover such eventualities.

Starting Protection

When a phase control circuit is first switched on it is quite possible that the thyristor firing angle will be fully advanced to give maximum power to the load. This will certainly be the case with simple furnace and motor control systems. A cold furnace or a stationary motor will demand full power from the control system. If the initial load impedance is very low, as with a shunt wound motor, the initial current when the supply circuit is made can be very large. No current limit circuit can operate until after the first half cycle, as there must be current before it can be used for control. In practice, several half cycles will be required before the current limit circuit can provide complete stable protection. During this interval it is quite possible that the control devices or the load could be damaged.

One way of guarding against such a condition is to arrange that, when the mains are first switched on, the thyristor firing angle is always fully retarded and is then slowly advanced at a slow enough rate to allow the current limit circuit to take over control. Such an automatic starting-surge suppression circuit must automatically reset whenever the mains circuit is broken if the protection system is to be foolproof. There are a number of ways in which such protection can be provided. The circuits of two systems are shown in Figs. 23 and 24. These circuits can be combined with the current limit circuits to provide full current limit protection during starting and running.

Fig. 23—Starting-surge protection circuit



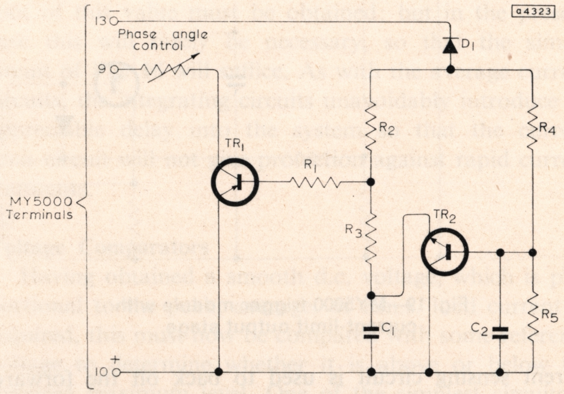


Fig. 24—Starting-surge protection circuit

The circuit of Fig. 23 shows a very simple starting surge suppression system. The resistor R_T is a small bead type thermistor with a relatively fast heating and cooling rate. Its resistance when cold is sufficiently high to fully retard the thyristor firing angle; but when heated by the current through it, its resistance falls to a low value so that it has little effect upon the firing angle setting. The resistor R_1 may be necessary to bypass additional current through the thermistor. This current heats the thermistor to a temperature where its resistance is so low that changes in ambient temperature have little effect upon the thyristor trigger circuit timing.

When the power supply to the trigger circuit is interrupted the thermistor cools, thereby resetting the surge suppression system. Provided that the surge suppression circuit is combined with a current limit circuit which has a fast charge time-constant and a discharge or reset time-constant longer than the cooling time-constant of the thermistor, the combined circuit will provide complete protection under all normal conditions of starting and running.

The circuit of Fig. 24 has much the same characteristics as the thermistor circuit. The transistor TR_1 is connected in series with the phase control circuit in the same way as the current limit output transistor, and under normal operating conditions is held in the bottomed state by the resistor chain R_1, R_2, and diode D_1. When the supply to the circuit is first made, C_1 is discharged, and TR_1 is held off while C_1 charges through R_2, R_3 and D_1 at a rate dependent on the time-constant C_1 (R_2+R_3). This time-constant is chosen to be several times greater than the charging time-constant of the associated current limit circuit. When the power supply to the circuit is interrupted, transistor TR_2 is turned on, discharging C_1, in order to reset the circuit. The small time-constant C_2R_5 is necessary to prevent the circuit resetting at each half cycle when the power supply to it is derived from a clipped but unsmoothed full-wave rectified supply, as used for the MY5000 trigger module. The diode D_1 prevents any interaction between the surge suppression circuit and the trigger circuit.

The circuit of Fig. 23 has the virtue of extreme simplicity and is therefore cheap and reliable; it however suffers from the disadvantage that the thermistor is in series with the timing circuit and therefore introduces a certain degree of firing angle drift with ambient temperature change. The circuit of Fig. 24 is largely free from this defect, but the change-over from surge suppression control to current limit control is not so smooth, a certain amount of overshoot seems to be inevitable, and the resetting time may be longer.

SOME PRACTICAL CIRCUITS

A simple current limit and surge suppression circuit is shown in Fig. 25, which is a combination of the circuits shown in Figs. 9, 12, 18, and 23. With the components shown the onset of current limit action corresponds with a voltage on the input capacitor of about 2·5V. This will be produced by a peak voltage of about 3V at the current transformer terminals, or an output voltage of about 2V r.m.s. with a sinusoidal input current. The circuit is stable for a load impedance range of at least 40:1, which should cover most normal requirements. Typical control characteristics are shown in Fig. 26 for a resistive load, from which it may be seen that the peak current increases slowly over the control range by a factor of a little less than 2 times. Because of the change in waveform over the range, the r.m.s. and average currents fall at first and then slowly rise, but never reach the maximum values again.

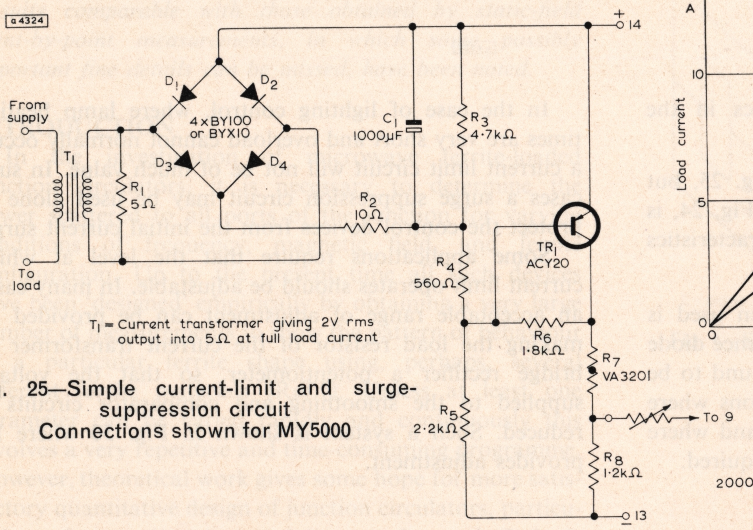


Fig. 25—Simple current-limit and surge-suppression circuit
Connections shown for MY5000

The effect upon the characteristics of varying R_6 is shown in Fig. 27. Increasing R_6 increases the gain of the feedback loop and improves the current control. If, however, R_6 is made too large the complete system may oscillate when the load resistance falls to some critical value.

When the current limit circuit is used in association with a d.c. shunt motor the current limit characteristics are modified somewhat by the fact that the motor is inductive, and a flywheel diode is used. The current waveform departs radically from that produced by a resistive load so that the relationship between peak and average current is quite different. In this case the characteristics are typically as shown in Fig. 28, where it will be seen that the average current increases slowly with increase of load torque and decrease of speed. The torque/speed curve

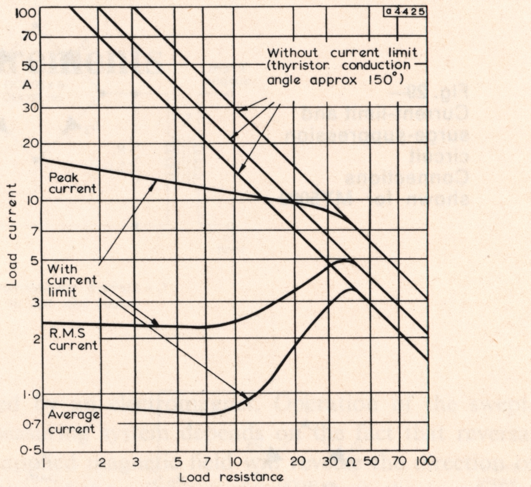


Fig. 26—Typical control characteristics of current-limit circuit

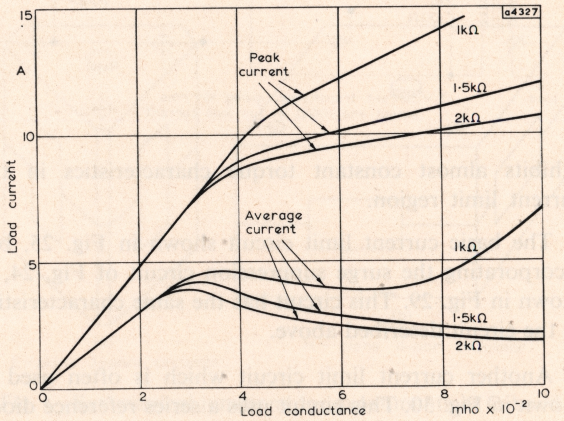


Fig. 27—Effect of variations in resistor R_6 in circuit of Fig. 25

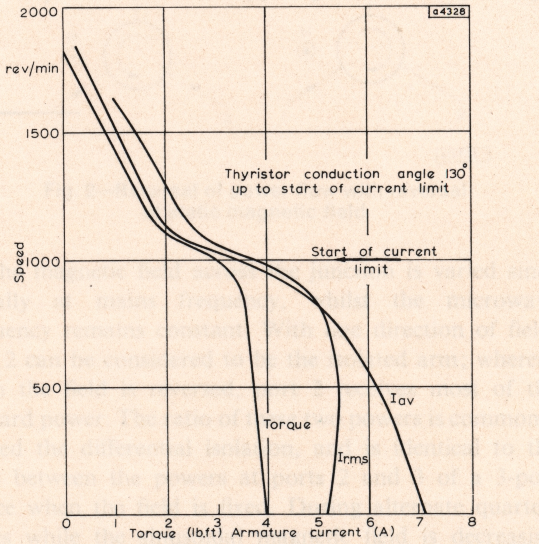


Fig. 28—Motor torque and current in current-limit region

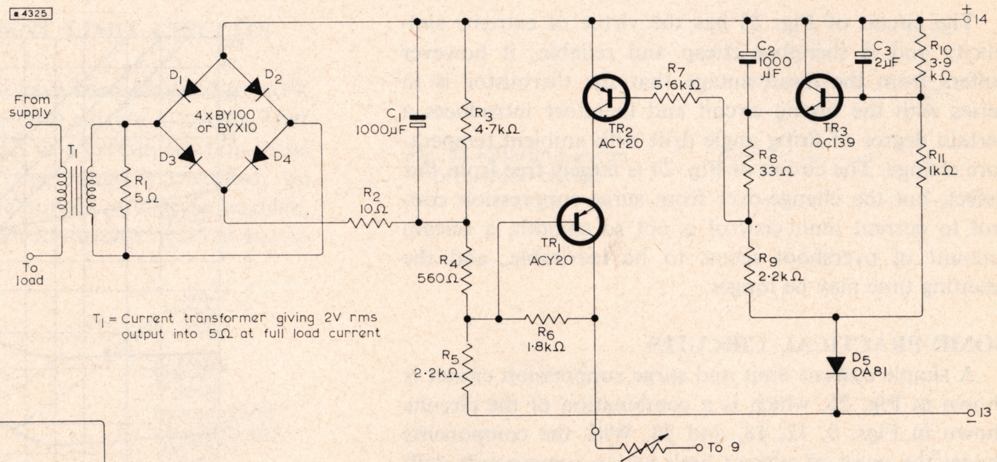


Fig. 29—Current-limit and surge-suppression circuit Connections shown for MY5000

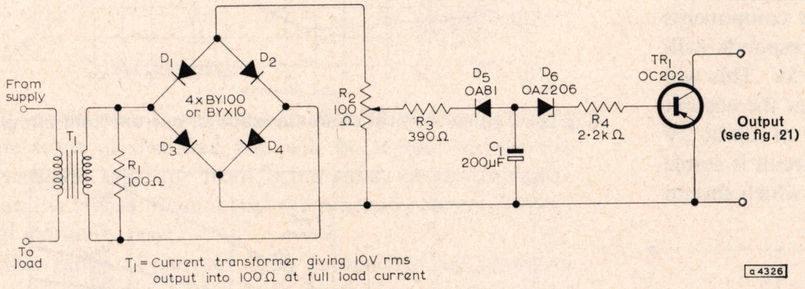


Fig. 30—Current-limit circuit

exhibits almost constant torque characteristics in the current limit region.

The basic current limit circuit shown in Fig. 25, but incorporating the surge suppression circuit of Fig. 24, is shown in Fig. 29. This circuit has the same characteristics as the circuit described above.

Another current limit circuit which is often used is shown in Fig. 30. This circuit uses a series reference diode and a simple output amplifier, and has been found to be satisfactory for many furnace control applications where the control range is limited to less than 10:1 and where rapid response and surge suppression are not required.

In the case of lighting control, where lamp heating times are very short and overload cannot normally occur, a current limit circuit will not be of much value. In such cases a surge suppression circuit may be used alone to protect the control devices from the initial current surge.

Some applications require that the level at which current limit operates should be adjustable. In many cases an acceptable range of adjustment can be provided by making the load resistor of the current transformer or bridge rectifier a potentiometer, so that the voltage supplied to the smoothing and comparator circuits is reduced. Such a system is shown in Fig. 30 where R_2 provides adjustment.

Swept-field System for Rapid Measurements on Microwave Ferrite Devices

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A method is described whereby rapid and accurate measurements on microwave ferrite devices may be carried out. The system, which does not require broadband microwave components or a swept-frequency signal generator, has been applied to circulator development, and it is shown that there are advantages over the more usual swept-frequency techniques, particularly in the millimetre wavelength range. Results comparable with those obtained by static-field point-by-point measurements, in which some possibly important fine details can be missed, have been noted.

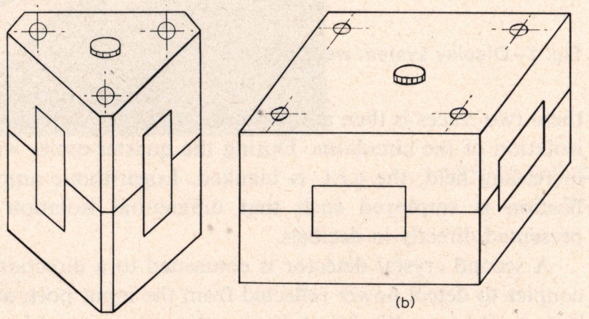
INTRODUCTION

In the investigation of microwave ferrite-loaded junction circulators, it is necessary to determine the power scattered to all ports of the junction for varying conditions of frequency, magnetic field, and ferrite configuration. Up to the present time all such devices have been 'designed' empirically by obtaining a very large number of measurements to show the pattern of behaviour of a particular ferrite shape using, perhaps, several different ferrite materials (Ref.1). Unless at least one parameter can be varied rapidly and continuously this involves a very repetitive and time-consuming programme. However, theoretical work gives some hope for more satisfactory quantitative design of junction circulators, particularly at frequencies where stripline can be used (Refs. 2, 3, and 4).

Since, in general, the behaviour of a circulator over a frequency band is of particular importance, swept-frequency measurements are desirable and give useful and rapid information. A description is given of a measuring system using a swept magnetic field. This was constructed for investigation of devices at 35Gc/s, but the principle provides a simple and inexpensive arrangement for work in any frequency band. An advantage is that standard waveguide components can be employed to give good accuracy. For example, very small reflection coefficients and insertion losses can be measured without bulky and expensive high-directivity broadband directional couplers and detectors. The information is recorded by presentation on an oscilloscope, the display being photographed with a Polaroid camera.

PRINCIPLE OF OPERATION

The magnetic field required for either H-plane or E-plane junction circulators is applied at right-angles to the plane of the junction, and for experimental work is provided by an electromagnet. Operation of the swept-field measuring system depends on the fact that reversal of the applied magnetic field will reverse the direction of circulation of the microwave power from the input port (Figs. 1 and 2).



(a) E-plane 3-port circulator

(b) H-plane 4-port circulator

Fig. 1—Symmetrical junctions for experiments at Q-band

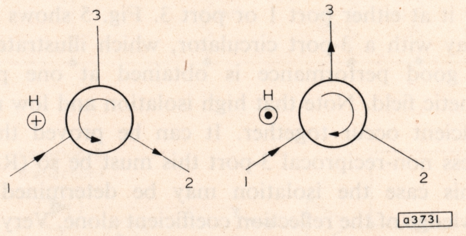


Fig. 2—Reversal of power flow with reversal of static magnetic field

The magnetic field across the junction is varied sinusoidally at mains frequency, whilst the microwave frequency remains constant. With one direction of field, port 2 can be considered to be the isolated arm; whereas when the field is reversed, port 2 receives most of the forward power. The ratio of these two powers is commonly termed the differential isolation, and is identical to the ratio between the powers at ports 2 and 3 of a 3-port device when the field is fixed. During alternate quarter-cycles while the sinusoidal magnetic field is decreasing from its maximum value to zero, the detected signal at port 2 is displayed on an oscilloscope. The horizontal timebase is driven by a full-wave rectified sinewave such that two traces appear on the screen, one showing isolation and the other insertion loss. The vertical distance between

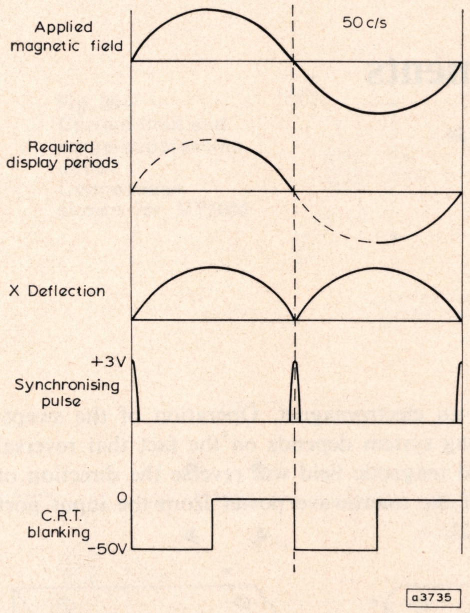


Fig. 3—Display system waveforms

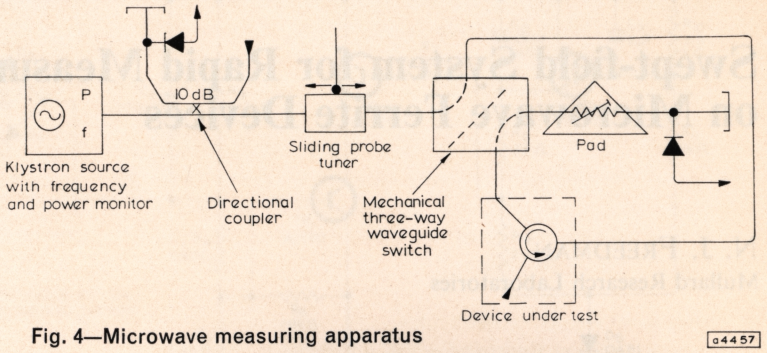

P
f
Klystron source with frequency and power monitor
10 dB
Directional coupler
Sliding probe tuner
Mechanical three-way waveguide switch
Pad
Device under test
a4457

Fig. 4—Microwave measuring apparatus

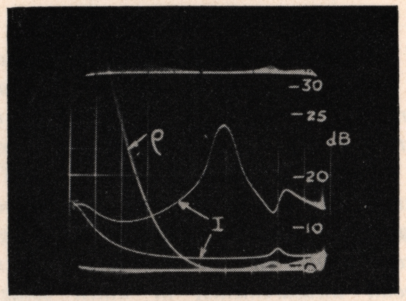


Fig. 5—Typical display of circulator isolation and reflection. Reflection coefficient photographed on same film by two exposures, using single c.r.t. beam. Signal modulated by square wave at 3kc/s

these two traces is then a direct measure of the differential isolation of the circulator. During the quarter-cycles with increasing field, the c.r.t. is blanked. Logarithmic amplification is employed such that differential isolation is presented directly in decibels.

A second crystal detector is connected to a directional coupler to detect power reflected from the input port, and its output is amplified and displayed on the second beam of a double-beam oscilloscope. Fig. 3 shows the relevant waveforms in the display section. In the microwave circuit depicted in Fig. 4, the 4-port 3-way mechanical waveguide switch allows the signal to bypass the circulator, or to enter it at either port 1 or port 3. Fig. 5 shows a typical display with a 3-port circulator, which illustrates clearly that good performance is obtained at one particular magnetic field. Note that high isolation and low reflection coefficient occur together. It can be proved that for a lossless non-reciprocal 3-port this must be so (Ref.5) and in this case the isolation may be determined from a knowledge of the reflection coefficient alone. Very accurate measurement of reflection coefficient is possible by tuning the directional coupler for perfect directivity at each operating frequency.

DISPLAY SYSTEM

A.C. Magnet

An electromagnet with a laminated yoke is essential if large fields are to be achieved. The usual iron-cored electromagnet used for point-by-point measurements would become heated, and the maximum field reduced, due to hysteresis and to eddy currents.

A large C-core, which may be seen in Fig. 6, was found to overcome these difficulties satisfactorily. Using a coil wound with 1300 turns of 20 s.w.g. wire and a d.c. coil current of 2A, a field of 3700 oersted was obtained in the 0·25in gap required for the H-plane junction circulator at 35Gc/s. For the E-plane junction with a gap of 0·36in, the field was 2400 oersted. In both cases the field varied linearly with current up to 1A, and the approach of saturation was becoming evident at 2A. A usable variation was, however, obtainable up to this point.

The coil had an inductance of 0·6H and was driven from the 50c/s mains via a variable transformer which provided a suitable control on the range of swept magnetic field.

X-deflection Waveform

The X-deflection waveform for the oscilloscope is obtained by full-wave rectification of the voltage appearing across a small resistor in series with the magnet coil.

An isolating transformer is necessary since it is desirable to earth the display circuits and the oscilloscope, and also the auto-transformer shown in Fig. 7.

The phase-shifting network is required to bring the X-deflection waveform into phase with the magnetic field in the air gap of the magnet. There exists a phase difference around the magnetic circuit between coil current and magnetic field, due to residual eddy currents and hysteresis. Setting can by done by means of a small search coil or a Hall-effect probe in the air gap, taking care to avoid direct pick-up from the magnet into the leads.

It is a simple matter to adapt the principle of the display for measurements on other non-reciprocal devices which depend on a parameter other than frequency. For example, Faraday-rotation devices can easily be investigated by deriving the c.r.t. X-deflection waveforms from the magnetising coil current as described.

Trace Blanking

To permit display of circulator parameters only during either 'upward' or 'downward' traverses of magnet and microwave ferrite hysteresis loops, the oscilloscope beam is blanked during alternate quarter-cycles of magnetic field. To obtain accurate timing, a sharp triggering waveform is derived to synchronise a 100c/s square-wave generator which supplies 50V to the c.r.t. grid. The polarity of this waveform determines the direction of magnetic field during which display occurs; that is, increasing or decreasing. The synchronising pulse is obtained by amplification, limiting, and differentiation of the X-deflection waveform shown in Fig. 3.

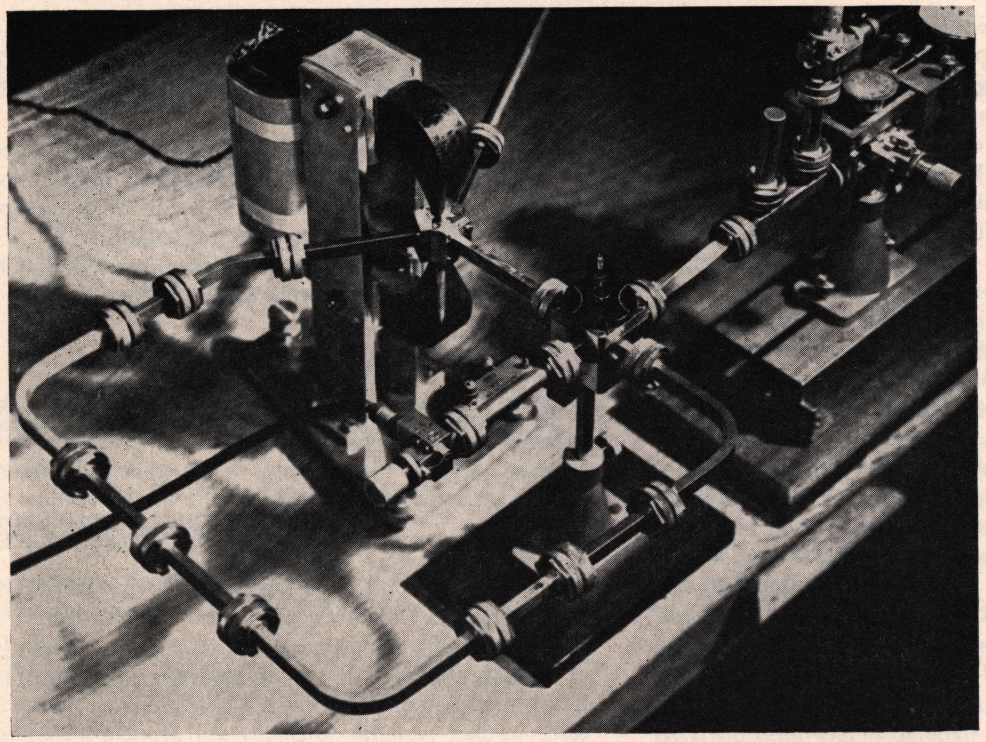


Fig. 6—Microwave measuring apparatus

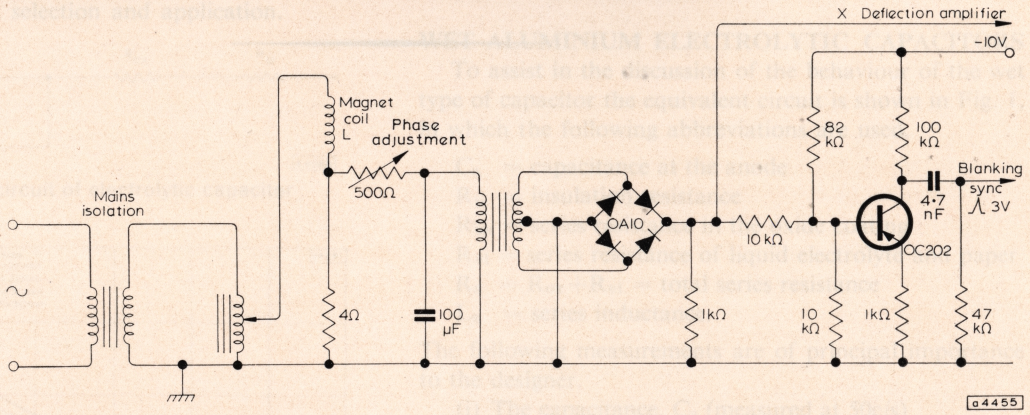


Fig. 7—Display circuit

Logarithmic Amplification

The dynamic range of isolation to be measured is usually about 35dB, therefore logarithmic amplification of the output from the signal detector is desirable if satisfactory resolution at both high and low isolations is to be available. High accuracy of isolation measurement is rarely required of a swept-parameter system during development stages of a circulator; a simple arrangement using a single semiconductor diode as the logarithmic element is therefore feasible (Fig. 8). The SX640 diode has a well-defined logarithmic law over eight decades of current, given by:

$$i = 10^{-10}(e^{23V} - 1)$$

which reduces to:

$$V = 1 + 0{\cdot}1 \log_{10} i$$

when V is greater than 0·1V.

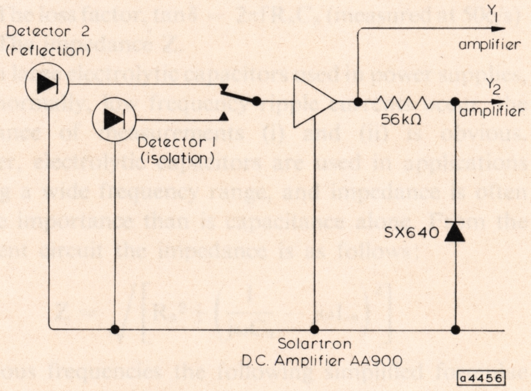


Fig. 8—Logarithmic circuit

The diode must be current-fed from a source impedance which is greater than the combined resistances of the diode and oscilloscope input in parallel.

The diode is then operated at as large a current as will give a good logarithmic characteristic, and the current driven by the largest available voltage via a high resistance. The Solartron d.c. decade amplifier type AA900 was found to be entirely suitable for the application. In practice, the d.c. facility has been found useful whenever static measurements are made, though with the swept system the lowest frequency to be amplified is 50c/s. With maximum microwave signal the amplifier gain is set to deliver about 25V into 56kΩ, when normal microwave crystal detector padding is used at Q-band; that is, when the v.s.w.r. is approximately 1·06.

The system is calibrated by switching the signal power directly to the detector, varying the level by measured amounts, and marking a scale on the oscilloscope screen. The calibration must always be carried out with a known maximum voltage output from the video amplifier into the logarithmic circuit. This level is continuously monitored; when a change of microwave frequency is made, padding must be adjusted to restore this voltage at the '0dB' level after each trace calibration. The level will, of course, differ for c.w. and for modulated signals if, for example, an Avometer is used to monitor the amplifier output.

It is preferable to view reflection coefficient on a linear scale, and to have the facility for displaying insertion loss also on a linear scale when it is small. With suitable junctions of leads this can easily be arranged.

The logarithmic amplification circuit used is equally applicable to swept-frequency measurements.

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Service-life and Reliability of Mullard Electrolytic Capacitors

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The construction and characteristics of different types of electrolytic capacitors are discussed and related to their service-life and reliability. Aid is given to the designer in the selection of the most suitable electrolytic capacitor for any particular application.

INTRODUCTION

Information dealing with the service-life and reliability of all capacitors is becoming increasingly important. It is estimated that one-quarter of all components used in electronic equipment are capacitors. One-seventh of all equipment failures are capacitor failures; half of these are caused by incorrect selection and application.

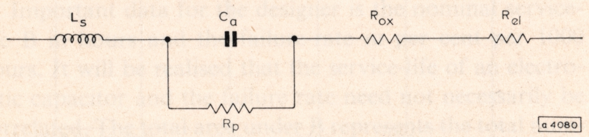

Fig. 1—Equivalent circuit of electrolytic capacitor

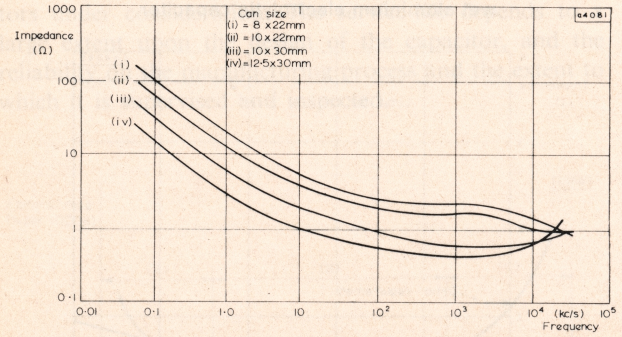

Fig. 2—Typical curve of impedance of wet aluminium electrolytic capacitors plotted against frequency

The transistorisation of electronic equipment has brought about large requirements for electrolytic capacitors. During the last decade a considerable amount of research has been carried out on this type of component, and as a result, major design improvements on existing components have taken place, and new types developed.

The general concept of the electrolytic capacitor, as unreliable and unpredictable in operation, needs to be revised. Electrolytic capacitors, including those of the wet aluminium type, can give extremely long service if properly manufactured and carefully selected for the application. Because of the multiplicity of types available and the complex behaviour of electrolytics, correct selection is not a straightforward task for the user. In this article factors affecting service-life and reliability are examined. This information, combined with the statistical data, will help the user to select the correct component, from a commercial as well as technical standpoint.

The article starts by examining the behaviour of wet aluminium electrolytic capacitors under different electrical and environmental conditions, which then makes it possible to state rejection criteria. This is followed by a general survey of the factors influencing service-life and reliability.

Solid electrolytic capacitors, which are becoming more widely used, have been treated separately because their electrical behaviour is different from that of the wet type.

WET ALUMINIUM ELECTROLYTIC CAPACITORS

To assist in the discussion of the behaviour of the wet type of capacitor the equivalent circuit is shown in Fig. 1, in which the following abbreviations are used.

C_a = capacitance at the anode
R_p = insulation resistance
R_{ox} = series resistance in the oxide coating
R_{el} = series resistance of liquid electrolyte and paper
R_s = $R_{ox}+R_{el}$ = total series resistance
L_s = series inductance.

The following measurements are of principal importance to the designer.

(i) The capacitance, C_a (measured at 50c/s)
(ii) The loss factor, $\tan\delta = 2\pi f R_s C_a$ (measured at 50c/s)
(iii) The impedance Z.

With large electrolytic capacitors used in power supplies, where normally, low frequency ripple currents occur, the significance of measurements (i) and (ii) is obvious. However, electrolytic capacitors are used in applications covering a wide frequency range, and impedance is often of more importance than is capacitance alone. From the equivalent circuit the impedance is as follows:

$$Z = \sqrt{\left[R_s^2+\left(\frac{1}{j\omega C_a} - j\omega L_s\right)^2\right]}$$

At various frequencies the following simplified formulae may be used:

$$\text{at 50c/s,} \qquad Z \simeq \frac{1}{\omega C_a}$$
$$\text{at 100kc/s,} \qquad Z \simeq R_{el}$$
$$\text{at 10Mc/s,} \qquad Z \simeq \omega L_s\,.$$

The above expressions show that, over a given frequency range, the impedance has a minimum value which is mainly presented by the liquid electrolyte. Fig. 2 shows a typical curve of impedance versus frequency.

The value of series resistance presented by the liquid electrolyte is important both as a parameter of service-life and as a factor which determines the behaviour of the component at low temperatures. The impedance measured at a frequency of 100kc/s is regarded as one of principal importance, since this approximates to the actual resistance.

Operation

When a d.c. voltage is applied to the component, a small current, called the leakage current, will flow. This value will be determined by the parallel resistance R_p. The dielectric of the capacitor, the oxide coating, is slightly attacked by the liquid, but is re-formed by the electrolytic action occurring within the capacitor when in use. Gases are released by the electrolytic action and diffuse outwards through the sealing of the capacitor. The liquid electrolyte becomes exhausted as a result of this process and a gradual increase in the resistivity of the liquid occurs. This process physically determines the life of the capacitor. The impedance and loss angle will increase and the effect of this on the capacitance is not obvious. It can be explained by consideration of the effect of the etched surface of the anode on the total capacitance. Fig. 3 gives a section through this surface. The total capacitance C_a is in fact made up of a number of capacitive components, where a component contained in a pore has a high series resistance in comparison with one on the surface. Fig. 4 shows a good approximation of such a situation. If now the liquid resistance increases, some of the capacitive components will lose much of their effect and this will show up as a drop in the total capacitance. Summarising, it can be said that the drying out of an electrolytic capacitor in service is shown by:

(i) A gradual decrease in capacitance (an increase in impedance at 50c/s).
(ii) An increase in losses (tanδ).
(iii) An increase in the impedance at high frequencies.

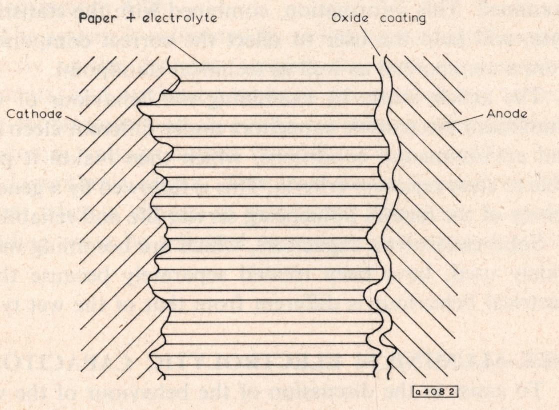

Fig. 3—Section through capacitive interface in electrolytic capacitor

Typical curves of the changes in capacitance and r.f. impedance of electrolytic capacitors during life tests are shown in Fig. 5 and Fig. 6.

Rejection Criteria

Now that the properties affecting the behaviour of the capacitor are known, we can establish failure criteria. First of all we must distinguish between two types of failure: these are catastrophic, and change failure. Catastrophic failures include open circuits, short circuits, and unpredictable and rapid departures from the normal characteristic values. Change failures occur when the electrical limits imposed on the component are exceeded as a result of gradual changes in the properties of the component during service-life.

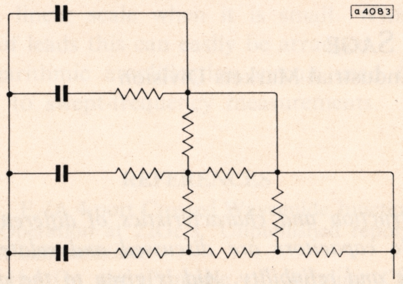

Fig. 4—Approximate form of equivalent circuit of capacitive interface in electrolytic capacitor

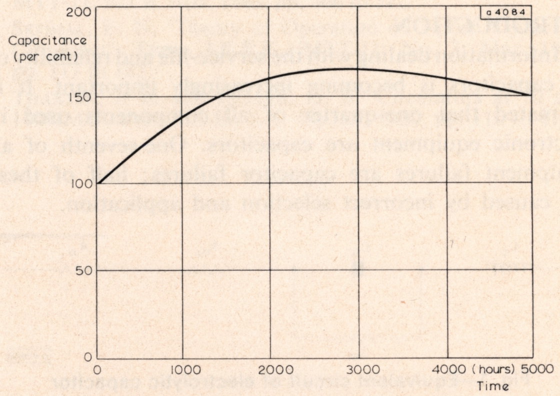

Fig. 5—Typical curve of capacitance against length of life for wet aluminium electrolytic capacitor

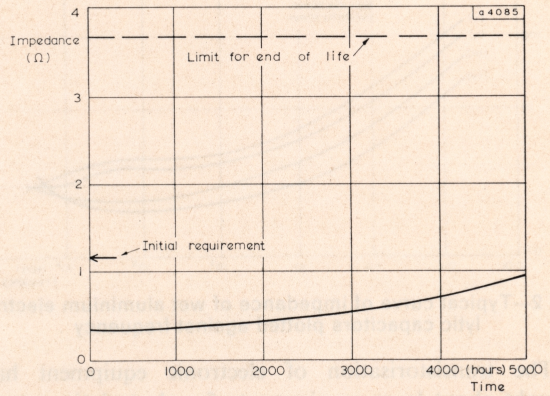

Fig. 6—Typical curve of impedance against length of life for wet aluminium electrolytic capacitor

A wet aluminium electrolytic capacitor will be regarded as having failed if one or more of the properties given below increases beyond the stated limits. (The figures in brackets apply to capacitors with a working voltage of less than 25V.)

(i) Impedance at 50c/s and at 20°C shall not exceed 150% (200%) of the initial value specified.
(ii) Dissipation factor at 50c/s and at 20°C shall not exceed 200% (300%) of the initial value specified.
(iii) R.F. impedance at 100kc/s and at 20°C shall not exceed 300% of the initial value specified.

SERVICE-LIFE AND RELIABILITY

If the designer has to choose an electrolytic capacitor from a large number of available types, some basic knowledge of the behaviour of such types and about the factors that influence their service-life and reliability is essential. These problems are discussed in detail below.

From a given batch of electrolytic capacitors, failures will arise in service. There is a fairly constant failure rate: this gives a guide to the reliability of the component. After a certain time, the gradual changes that take place in the properties of the capacitor reach a failure point evidenced by a sudden and marked increase in the total failure rate. The moment at which this occurs is a measure of the average life of the capacitor type in question. The curve in Fig. 7 gives failure rate for wet electrolytic capacitors as a function of time. Three distinct zones may be seen in Fig. 7; these are as follows:

(i) The failures during manufacture and final inspection.

(ii) The failures during service-life.

(iii) The failures at the end of useful life.

Important data for the designer is the nominal service-life B in hours and the failure-rate in per cent per 1000 hours. It will be realised that the service-life of an electrolytic capacitor and the failure rate need not necessarily be correlated. The total area under B represents the total number of failures during the whole of the useful life and is generally only a small percentage of the number of capacitors under consideration. The service-life depends to a large extent upon the design of the capacitor, and the reliability on the manufacturing process and the extent to which it is supervised and inspected.

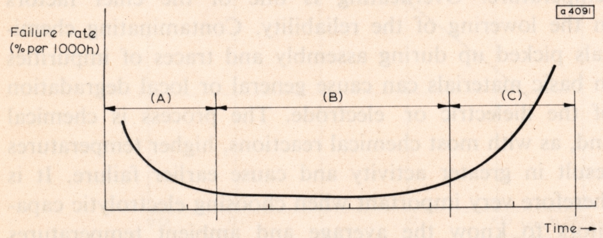

Fig. 7—Failure rate curve for wet electrolytic capacitors

Factors Affecting Service-life and Reliability

The factors affecting service-life and reliability fall into two categories which can be called internal and external factors. Internal factors are those determined exclusively by the design of the capacitor; the finished product is thus already intended for a range of applications because the designer has achieved the best compromise between the various relevant requirements. It is, however, possible in the case of electrolytic capacitors for the user to make the capacitor even more suitable for his specific purpose by careful choice and taking into consideration the external factors likely to affect the service-life and reliability.

Table 1 gives an outline of the most important factors which will be discussed in detail.

TABLE 1

Factors Affecting Service-life and Reliability

Class of Factor	Variable	Group Responsible
Internal	Purity of materials Choice of liquid electrolyte Sealing	Development laboratories
	Supervision of manufacture Final inspection Random testing	Factory
External	Temperature Voltage Dimensions	User

Although the user is not directly concerned with the internal factors, it may be useful for him to gain some insight into these points.

Purity of Materials

Impurities, particularly any chlorine ions, present in the materials used (that is, aluminium foil, paper, sealing discs etc.) can cause leakage currents resulting in the speeding up of the drying out process, and in the shortening of service-life. A prime requirement for making a reliable and high quality product, therefore, is the use of very pure materials and a continuous check on their quality.

Choice of Liquid Electrolyte

There are a very large number of chemical compositions used for liquid electrolytes. Two main requirements, which are unfortunately diametrically opposed, generally need to be observed. One is the need for low resistance at low temperatures, and the other is that there should be only a small increase in resistance during service. The choice of electrolyte is nearly always a compromise between these two requirements. Since the overall stability of the capacitor, and hence its service-life, is directly connected with the type of electrolyte selected, a major part of development work is carried out on this subject. Another important variable is the permissible working temperature. One of the inherent drawbacks of good performance at low temperature is less favourable performance at high temperatures. Here, too, a compromise has to be sought, and the requirements of the particular application envisaged will affect the choice. To illustrate these points Fig. 8 gives impedance versus temperature curves for three types of electrolytic capacitor (C421 is a solid type). The initial curves and curves after 1000h at full electrical loading are given. It is quite possible that the capacitor with the best performance at low temperatures could have a shorter service-life. Solid electrolytic capacitors are an exception to this. The manganese dioxide replacing the liquid electrolyte has a resistance that only slightly depends upon temperature and has good long-term stability.

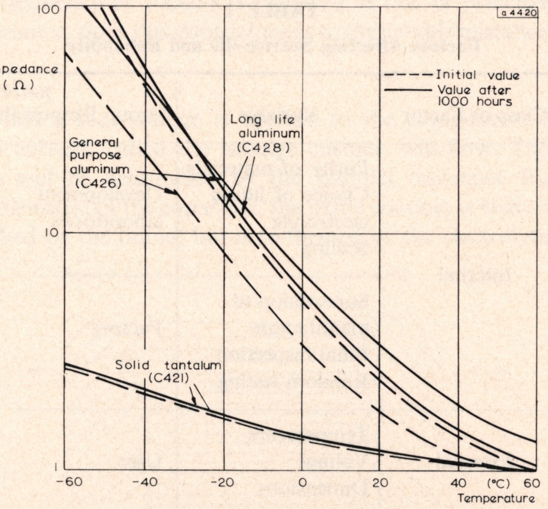

Fig. 8—Curves of impedance of various electrolytes plotted against temperature

Sealing

The sealing of a wet electrolytic capacitor is very important, since during the process of electrolysis the gases developed in the capacitor must be able to diffuse through the component. The capacitor must also be sufficiently well sealed to prevent the fluid electrolyte from leaking. Solid types, on the other hand, need to be well sealed against humidity, and since they give off no gas this requirement can be achieved.

Control of the Manufacturing Process

The best designed electrolytic capacitor will not turn out to be a good product unless the manufacturing process can be fully controlled and checked.

Final Inspection

Detecting and eliminating early stage failures (area A in Fig. 7) is a task that must be based on a final inspection which is 100% efficient in checking the main properties of the capacitor, such as leakage current, capacitance and r.f. impedance.

Random Sample Checks

It is impossible to maintain a high level of quality merely by the means already discussed. Independent sample tests must also be carried out. This sort of check, particularly in the case of electrolytic capacitors in which the initially measured values are not static, will be useful when made on samples taken at random from the production line. The main variables are controlled in this way.

Ambient Temperature

As previously stated, the leakage current of a wet electrolytic capacitor directly affects its life. Increasing the temperature causes the oxide coating to be more violently attacked, thus increasing the leakage current necessary for the 'healing' process. Fig. 9 gives a typical curve for the leakage current as a function of temperature.

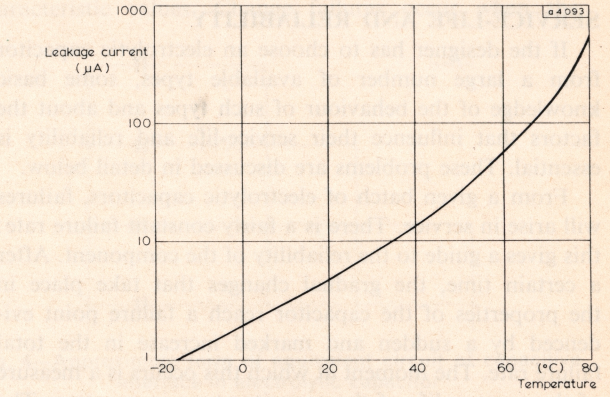

Fig. 9—Typical curve of leakage current against temperature, for electrolytic capacitor

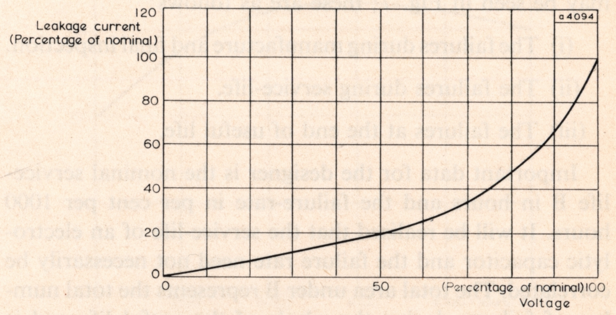

Fig. 10—Typical leakage current plotted against voltage for wet aluminium capacitor

It shows that after an initial gradual rise the increase becomes markedly progressive and can finally lead to the destruction of the capacitor.

Reliability, as well as service-life, is governed by temperature. Overheating is one of the chief factors in the lowering of the reliability. Contaminating chemicals picked up during assembly and traces of impurities in basic materials can cause general or local degradation of the dielectric or electrode. The process is chemical and, as with most chemical reactions, higher temperatures result in greater activity and cause earlier failure. It is therefore very important when choosing electrolytic capacitors to know the average and ambient temperatures at which they will be used and to take this into account when selecting the most suitable type.

Applied Voltage

It is well-known that the wet aluminium electrolytic capacitor is a polar component that may only be biased in one direction with a d.c., or a combined a.c. and d.c., voltage. The capacitor is designed for a particular working voltage, and at this loading it has an appropriate leakage current. Fig. 10 gives the leakage current as a function of the applied voltage. This current can be considerably reduced by under-loading the component. This results in an increased service-life.

The question often arises as to whether use at low voltages has disadvantages. The general rule is that an increase in capacitance is unavoidable, but this will seldom be regarded by the user as a drawback. It is important to remember that the stability of capacitance is closely bound up with the rated working voltage. Oxide coatings designed for working voltages of 25V, or more, are very stable and seldom cause increases in the capacitance of more than 10 to 15%. The thinner coatings, however, for lower voltages, tend to be less stable. It is, however, clear that these drawbacks may be acceptable with the improved service-life. It is worth noting that if long life is not a stringent requirement, operation at, or close to, the rated conditions results in minimum size, weight and cost. The effect of voltage derating upon the reliability of wet types is much less pronounced than the effect of changes in the ambient temperature.

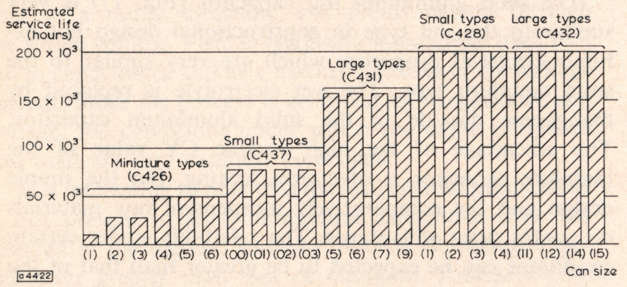

Fig. 11—Service-life shown against capacitor dimensions

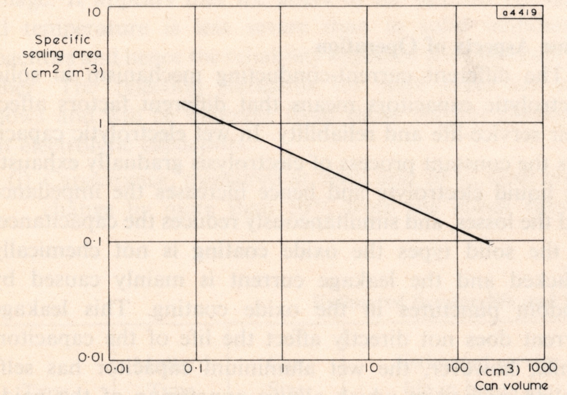

Fig. 12—Specific sealing area plotted against volume of can

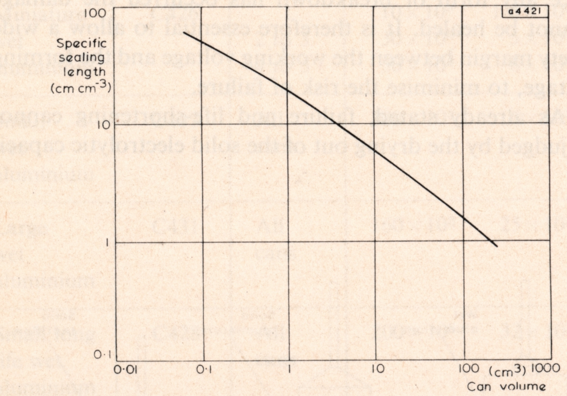

Fig. 13—Specific sealing length plotted against volume of can

Dimensions

Development of existing techniques and the introduction of new ones, has led to a marked reduction in the size of electrolytic capacitors. There is, however, the question as to whether this reduction is compatible with long service-life and outstanding reliability. Such compatibility does not appear to be the case if we look at Fig. 11 which shows estimated service-life as a function of the dimensions of the capacitor.

To explain this phenomenon, we must look at several aspects of manufacture more closely. Two specific cases are the problem of sealing small capacitors, and contamination by chlorine ions. The area and circumference of the seal must be considered in relation to the volume of the capacitor. The smaller this specific sealing length and area, the better is the seal and the less risk there is of quick drying-out and of leakage. Relationships between these factors and the volume of the capacitor may be seen in Figs. 12 and 13. The larger the can size, the more effective the sealing.

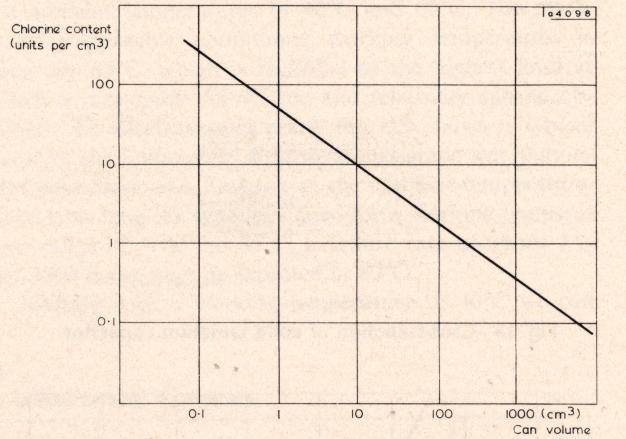

Fig. 14—Specific chlorine ion content plotted against can volume

It has already been mentioned that contamination by chlorine ions has a very marked effect on the leakage current and therefore on the service-life; such impurities may be introduced during the manufacturing process. If we consider that similar processes and materials are employed for both small and large capacitors, the quantity of impurities introduced per unit volume is considerably greater for the smaller types. This is illustrated in Fig. 14, which gives a specific chlorine content as a function of the can volume.

The external factors affecting service-life and reliability of wet aluminium electrolytic capacitors lead to the following general 'rules'.

(i) Operate the component within its temperature range, and even then at the lowest possible maximum temperature.
(ii) Voltage derating can be used as an important means of extending the service-life.
(iii) Use the largest possible can-size.

SOLID ELECTROLYTIC CAPACITORS

The development of solid types of electrolytics was prompted mainly by the limitations of the wet electrolytic types under some severe working conditions. The most important capacitors in this group are the solid tantalum slug capacitor, and solid aluminium foil capacitor.

The solid tantalum type is the most widely known, and is quite different in design from the orthodox wet aluminium foil variety. The anode is a pressed and sintered tantalum powder slug, coated with an oxide layer which forms the dielectric. A solid conducting coat of manganese dioxide acts as the electron conductor and replaces the ionic conduction of the liquid electrolyte in wet types. A layer of graphite forms the connection with a silver or copper layer, the whole being soldered in a hermetically sealed can. Fig. 15 shows a section through the capacitor. The conductivity of the manganese dioxide is governed only to a slight extent by temperature. Since the impedance at frequencies above 10kc/s is determined mainly by the series resistance in the manganese dioxide layer, its curve as a function of temperature is extremely 'flat'. Fig. 16 gives a typical example of this curve, showing the impedance curves of both a solid tantalum and wet electrolytic capacitor.

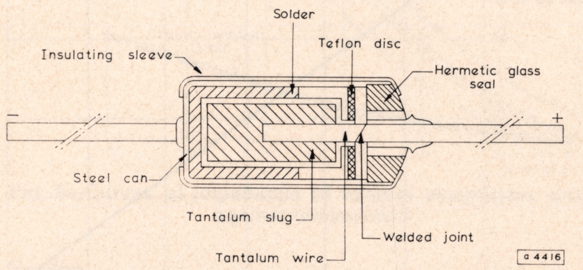

Fig. 15—Cross-section of solid tantalum capacitor

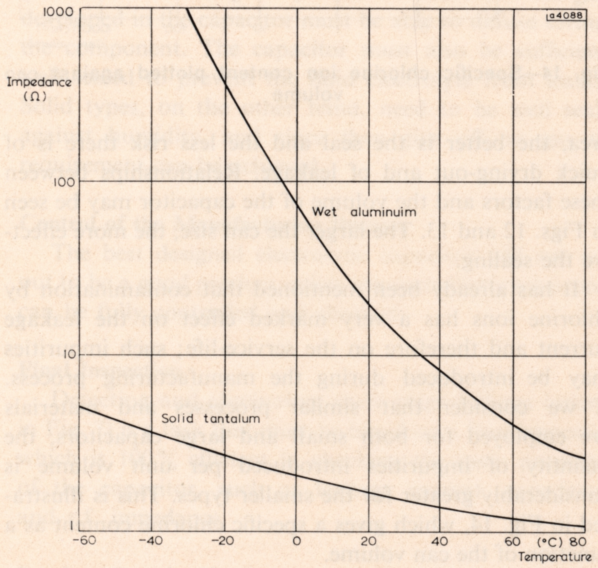

Fig. 16—Impedances of wet aluminium and solid tantalum capacitors, plotted against temperature

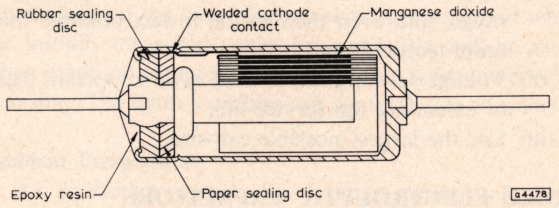

Fig. 17—Cross-section of solid aluminium capacitor

The solid aluminium foil capacitor (Fig. 17) is very similar to the wet type in constructional design but exhibits electrical properties which are very similar to the solid tantalum type. The wet electrolyte is replaced by manganese dioxide in the solid aluminium capacitor. The advantage of the high specific CV value of the tantalum capacitor is of course lacking, but the simple design and the comparatively inexpensive raw materials do confer price advantages. The reliability under certain conditions can be expected to be greater than that of the solid tantalum type. This point is more fully discussed later in the article. Fig. 18 shows comparative sizes for solid and wet types of capacitor.

Some Aspects of Operation

The different current-conducting mechanism in solid electrolytic capacitors means that different factors affect their service-life and reliability. In wet electrolytic capacitors the constant process of electrolysis gradually exhausts the liquid electrolyte and hence increases the impedance and the losses, and simultaneously reduces the capacitance. In the solid types the oxide coating is not chemically attacked and the leakage current is mainly caused by random punctures in the oxide coating. This leakage current does not directly affect the life of the capacitor. While, however, the wet aluminium capacitor has self-healing properties which allows puncturing of the oxide coating to be made good, this is not so with solid types: once any form of breakdown has occurred the damage cannot be healed. It is therefore essential to allow a wide safety margin between the working voltage and the forming voltage, to minimise the risk of failure.

As already stated, failure and life-shortening cannot be judged by the drying out of the solid electrolytic capacitor. One of the principal phenomena leading to failure in a solid tantalum capacitor is the chemical transformation of the oxide coating, known as field crystallisation. The dielectric normally consists of an amorphous oxide, but local growths of crystal can occur. These start at the surface of the tantalum and then spread through the amorphous coat in a radial direction, and finally extend through the whole volume. Field crystallisation manifests itself as a fall in capacitance and as a marked increase in the leakage current. Since the process is governed by temperature and voltage, the conditions under which the oxide is formed and used are very carefully controlled.

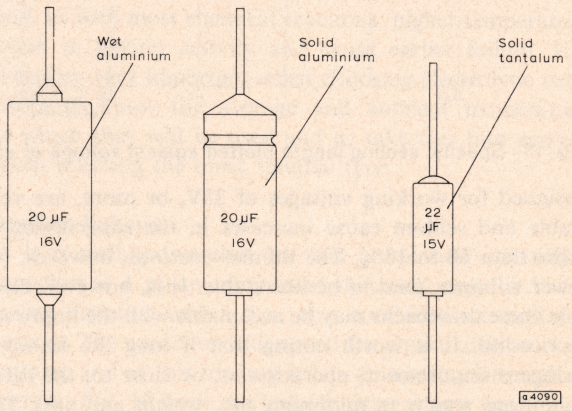

Fig. 18—Comparative sizes of various types of capacitor

It is interesting to note that the field crystallisation effect is not present in solid aluminium capacitors. This has resulted in the thermal stability of solid aluminium types being far greater than that of tantalum. Extensive tests have proved that the solid aluminium capacitor is capable of standing reverse and a.c. voltages without permanent damage. It appears that the effect of the applied voltage and temperature is less severe than in solid tantalum capacitors and hence the reliability expected is considerably higher. The rejection standards for solid types are as follows and will be different from those for wet aluminium electrolytic capacitors.

(i) Capacitance, measured at 50c/s and 20°C, shall not change by more than 10% from the initial value specified.

(ii) Dissipation factor, measured at 50c/s and 20°C, shall not exceed the initial value specified.

(iii) Leakage current, measured at 20°C, shall not exceed 200% of the initial value specified.

(iv) R.F. impedance, measured at 100kc/s and 20°C, shall not exceed 120% of the initial value specified.

FAILURE RATE AND SERVICE-LIFE DATA

Service-life and failure rate data are given in Table 2 below. These figures apply at maximum voltage and continuous service.

Wet Aluminium Electrolytic Capacitors

The service-life and failure rate of wet types are given for ambient temperatures of 40°C and 60°C. The maximum permissible continuous working temperature is generally 85°C, which is regarded as the highest level at which a reasonable service-life and reliability can be obtained. To illustrate this point Fig. 19 shows a typical curve of the failure rate of professional-grade wet aluminium capacitors as a function of the ambient temperature. Even with long-life types designed for a working temperature range of −40° to 85°C, a failure rate of about 1% per 1000 hours may be expected at 80°C.

Starting with a working temperature of 40°C we can

TABLE 2
Estimated Service-life and Reliability of Electrolytic Capacitors

Type	Mullard Type No.	Can Size	Estimated service life in hours at:		Correction factor $K_{0\cdot5}$	Estimated failure rate in % per 1000h. at:			Correction factor $K_{0\cdot5}$
			40°C	60°C		40°C	60°C	80°C	
Miniature wet aluminium	C426	1 2 to 3 4 to 5	10×10^3 30×10^3 50×10^3	$1\cdot6\times10^3$ 5×10^3 8×10^3	3·0 2·5 2·0	0·7 0·3 0·2	4·5 2·0 1·2	— — —	0·6 0·6 0·6
Small wet aluminium	C437	All cans	100×10^3	16×10^3	1·5	0.16	1·0	—	0·8
Large wet aluminium	C431	All cans	160×10^3	25×10^3	1·2	0·1	0·6	—	0·8
Small long life wet aluminium	C428	All cans	200×10^3	32×10^3	1·0	0·025	0·16	—	1·0
Large long life wet aluminium	C432	All cans	200×10^3	32×10^3	1·0	0·04	0·25	—	1·0
Solid tantalum	C421	All cans	—	—	—	0·28	0·42	0·63	0·1
Solid aluminium	C415	All cans	—	—	—	0·1	0·15	0·25	0·1

Notes: (i) $K_{0\cdot5}$ = correction factor for operation at 50% of rated voltage
(ii) Service life figures for solid electrolyte tantalum and aluminium capacitors are not given, as the failure rate levels remain constant over a long period.

give a general rule for the effect of the temperature on the service-life. Any chemical reaction is accelerated by a factor of from 2 to 3 if the temperature is increased by 10°C. This leads to a factor of about 6, for the range 40 to 60°C. This factor increases the failure-rate figure and decreases the service-life. The effect of the applied voltage on service-life and failure rate is given in the data.

Solid Electrolytic Capacitors

The service-life of solid electrolytic capacitors is in theory infinitely long. Unlike wet types, solid electrolytic capacitors have a failure rate very dependent on the applied voltage. As can be seen from Table 2, solid tantalum capacitors have the same reliability as standard wet types at 40°C, although operation at half voltage gives a considerable improvement. Fig. 20 shows typical figures of failure rate as a function of temperature for different load conditions. A considerable improvement at higher temperatures than 40°C can be seen when compared with wet types. Fig. 21 shows typical failure rates for solid tantalum capacitors, as a function of applied voltage at different working temperatures.

There is another factor, apart from temperature and voltage, that affects the reliability of solid tantalum types. The failure rate will vary according to the value of series resistance between the capacitor and power source. The values given in the data apply with a circuit resistance of 3Ω/V. For lower values of series resistance a correction factor should be applied to the failure rate figures as given in Table 3. These correction factors for the failure rate are also shown in graph form in Fig. 22. The curve permits interpolation, and emphasises the increase in failure rate as series resistance approaches zero.

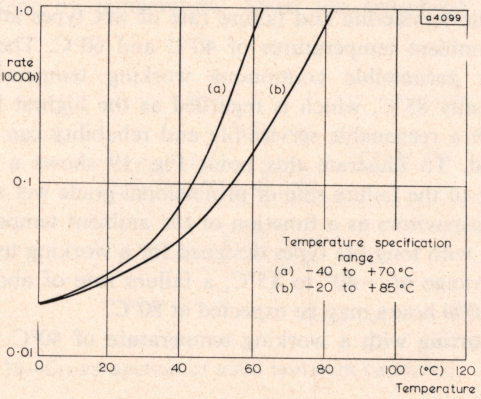

Fig. 19—Failure rate of professional-grade wet aluminium capacitors plotted against temperature

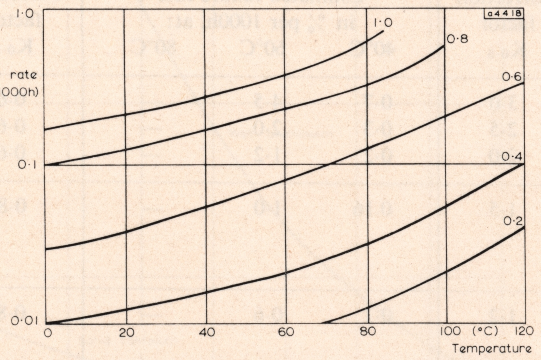

Fig. 20—Failure rate of solid tantalum capacitors plotted against temperature. Curves are labelled with voltage stress ratio

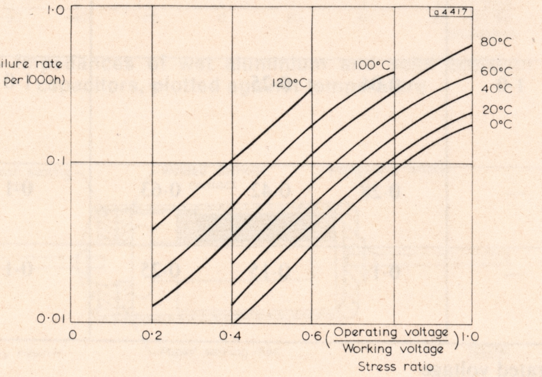

Fig. 21—Failure rate of solid tantalum capacitors plotted against applied voltage

TABLE 3
Correction Factor for Failure Rate

Circuit resistance (Ω/V)	Correction factor
3·0	1·0
2·0	1·4
1·0	2·7
0·9	2·9
0·8	3·2
0·7	3·6
0·6	4·0
0·5	4·5
0·4	5·1
0·3	6·1
0·2	7·5
0·15	9·0
0·1	12·0

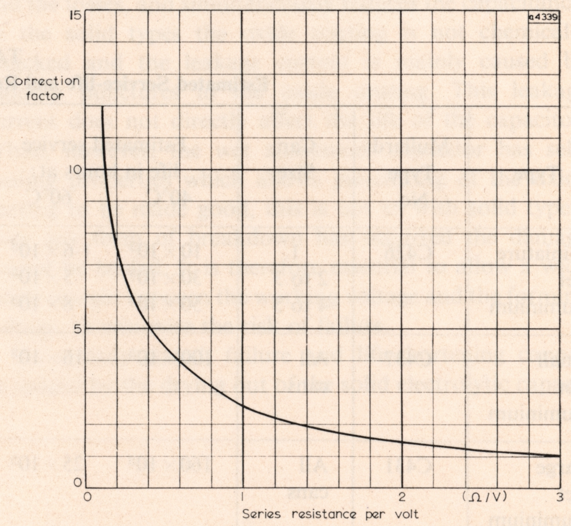

Fig. 22—Failure rate correction factor against series resistance

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Regenerative Braking of D.C. Motors

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Mullard Semiconductor Electrical Development
Laboratory and Central Application Laboratory

Of the four methods of braking d.c. motors, only one – regenerative braking – avoids the wasteful conversion of kinetic energy into heat. This method, in which energy is fed back into the supply during braking, requires special circuit arrangements if the d.c. supply is derived from a.c. mains via a thyristor bridge. The question is discussed in terms of operation from single-phase a.c. supplies, and then in terms of the more complex, but more usual, case of three-phase supplies.

An expression can be derived from which the regeneration current and motoring current can be calculated in terms of the circuit constants, the supply voltage, and the back e.m.f. The expression is identical to that derived by Vedder and Puchlowski (1943) for thyratron-equipped systems.

The principles discussed apply not only to motor braking but to any system in which energy is to be transferred from a d.c. source to an a.c. supply.

A practical control system, incorporating regenerative braking, is described for a d.c. motor operated from a three-phase bridge.

Principles of Regenerative Braking

D.C. machines are often used in applications which require energy to be fed back from the load into the supply during braking. A few examples of this are lifts, hoists, cranes, etc. The mechanism involved in this feedback process (usually called regeneration) is simple for d.c. supplies, but becomes rather complex when the d.c. supply is derived from a controlled rectifier bridge. This article describes in detail how regeneration can be achieved in such cases.

METHODS OF BRAKING

There are four methods of braking d.c. motors:

- (i) Mechanical
- (ii) Dynamic
- (iii) Plugging
- (iv) Regeneration

Mechanical Braking

Mechanical braking is the application of an extra mechanical load by means of band brakes or something similar.

Dynamic Braking

Dynamic braking is simply the application of an electrical load in the form of a resistor across the armature. Provided that the field is still excited, the motor will act as a generator, and the kinetic energy in the motor and load will be dissipated in the resistor and lost as heat.

Plugging

Plugging, in this sense, means the reversal of the d.c. supply to the motor. This applies a reverse torque and therefore brings the armature either to a standstill or to some speed in the opposite direction of rotation. This method is not often used on any but the smallest motors because of the very high currents that flow.

Regeneration

If the generated back e.m.f. of a motor is allowed to force current back into the supply, against the supply voltage, then energy is transferred from the motor to the supply. This is termed regenerative braking, because the total kinetic energy of the system (excluding losses) is returned to the supply by the motor, and hence it slows down. This is the only method of braking which does not lose the kinetic energy in the motor as heat. It is therefore the most efficient method of braking, and its use is essential in large motor control systems.

REGENERATION PRINCIPLE

The object of this article is to give, in the simplest possible terms, an outline of the operation of regeneration. Although the description refers to regeneration of d.c. motors, it should be appreciated that the principles apply to any system which requires the transfer of energy from a d.c. source to an a.c. supply.

The simplest form of regeneration is obtained when a d.c. motor is connected to a d.c. supply, as shown in Fig. 1.

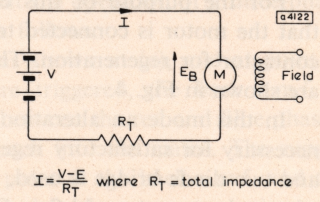


Fig. 1—Regeneration in d.c. system

When the d.c. supply voltage is greater than the back e.m.f. of the motor, then energy is taken from the supply to drive the motor. If the back e.m.f. rises above the supply voltage (that is, if the motor is accelerated to a higher speed by external mechanical means) then energy is transferred from the motor to the supply.

REGENERATION ON SINGLE-PHASE SUPPLIES

In a d.c. system it is easy to transfer the excess energy from the motor to the supply; but if the d.c. supply is obtained by rectifying the a.c. mains, then the operation of regeneration is more complex. This is because the energy must somehow be made to pass back through the rectifier into the a.c. mains.

Although regeneration is rarely required on single-phase mains, the single-phase system is simple to understand, and it provides a suitable introduction to the more complex system used on three-phase supplies.

Fig. 2 shows a d.c. motor connected to a single-phase supply via a half-controlled bridge. The conduction angle of the thyristors determines the mean value of the current through the motor and hence its speed for a given torque. Under these conditions (motoring) the back e.m.f. of the motor is less than the mean level of the full-wave rectified supply, and the voltage and current waveforms are as shown in Fig. 3.

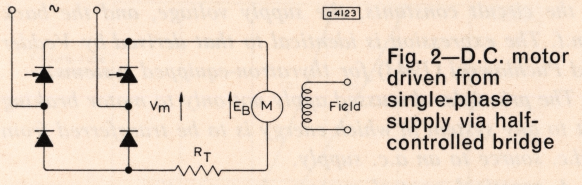


Fig. 2—D.C. motor driven from single-phase supply via half-controlled bridge

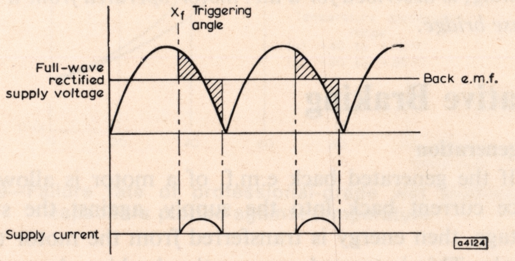


Fig. 3—Waveforms for Fig. 2 circuit during motoring

If the conduction angle is decreased, or the motor speed is increased by external mechanical means so that the back e.m.f. is greater than the mean output voltage of the bridge, the mean current would reverse if allowed. However, no regeneration can take place because current cannot flow in a reverse direction through the diodes or thyristors. Therefore the back e.m.f. of the motor must be reversed before current can flow from the motor to the supply.

For the purpose of this explanation we can assume that the motor is connected to the bridge via a reversing contactor for regeneration. The regeneration connections are shown in Fig. 4.

In this mode an alteration to the bridge of Fig. 2 is necessary for satisfactory regeneration. If a two-thyristor and two-diode bridge is used, and one of the thyristors is triggered, current could flow from the motor through the thyristor and diode back to the motor. The value of this current would be limited only by the resistance of the motor. Therefore four thyristors are used, and they are triggered in diagonal pairs.

The pairs of thyristors must be triggered in synchronism with the a.c. supply, and in such a way that the supply voltage is opposing the return of the current. In greater detail, the operation is as follows.

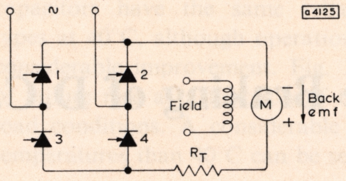


Fig. 4—Adaptation of Fig. 2 circuit for regeneration

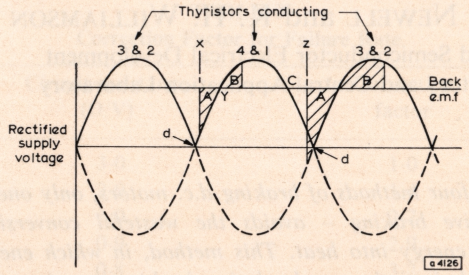


Fig. 5—Voltage waveforms for Fig. 4 circuit during regeneration

Let thyristors 4 and 1 (Fig. 4) be triggered at point x (Fig. 5). At this instant the back e.m.f. is greater than the supply voltage, therefore current will flow from the motor back into the supply. However, the supply voltage is increasing, and at point y it equals the back e.m.f. of the motor. Because of the inductance of the armature, the current continues to flow although the supply voltage becomes greater than the back e.m.f. The current ceases when area B equals area A. This is the discontinuous current condition.

It should be noted that the current flows against the supply voltage. That is, conventional current is flowing into the then positive terminal of the supply, which means that the supply is receiving energy from the motor.

If the back e.m.f. is increased so that area A increases, then, again because of the armature inductance, area B increases and current can be made to flow over the full period. This is the continuous current condition.

The current increases when the back e.m.f. is greater than the supply voltage, and decreases when it is smaller. Therefore if area B extends past point C, the current starts to rise again and continues to do so while the supply reverses and reaches a peak of the opposite polarity. At this point we have the back e.m.f. aiding the supply voltage; and, as the circuit resistance is very low, a large destructive current flows round the circuit. This is called 'plugging', and it must be avoided. The way to do this is to commutate the current from one pair of thyristors to the other pair just before the supply voltage reverses, so that the regenerative current is flowing against an opposing supply voltage for the major part of the cycle.

This is done by triggering the thyristors at point z just before the end of the half cycle. Assume thyristors 4 and 1 are conducting up to point z, and then at point z thyristors 3 and 2 are triggered. The polarity of the supply voltage at this instant is such (Fig. 5) that when thyristors 3 and 2 are triggered, thyristors 4 and 1 become reverse biased and turn off. Therefore, provided that during regeneration the trigger pulses to the thyristors always occur before point d, commutation of the current will take place and 'plugging' will be prevented.

As the back e.m.f. decreases, the regeneration current falls, so that point z can be advanced to allow more current to flow; and as point z is advanced, the motor goes from regeneration in one direction of rotation, stops, then to motoring in the opposite direction. If the amount of regeneration and acceleration is controlled by current limitation, then the motor can be made to reverse in the minimum safest time.

REGENERATION IN THREE-PHASE SUPPLIES

A three-phase system is very similar in operation and circuit layout to the single-phase system, therefore we can apply the same reasoning.

Consider the three-phase circuit of Fig. 6 and the three-phase waveforms of Fig. 7. The motoring condition waveforms are as shown in Fig. 7. The thyristors are triggered at points S, and the inductance of the motor armature causes the current to flow until the areas P and Q are equal. If the motor is loaded, the back e.m.f. falls, causing more current to flow.

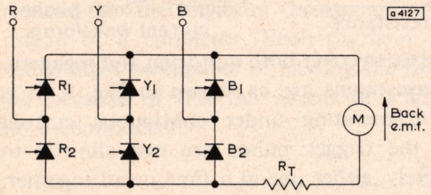


Fig. 6—Three-phase circuit during motoring

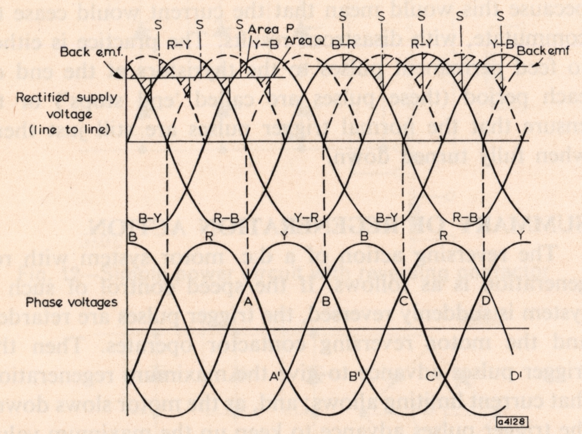


Fig. 7—Three-phase voltage waveforms during motoring

Before current can flow two thyristors must be triggered. The order of triggering is as follows. At point A the red phase and the yellow phase are of the same polarity and voltage, and at any point before A (up to 180°) the red phase is more positive than the yellow phase. So allowing the red phase to conduct into the yellow phase from this point will give smooth control over the output voltage. If the red phase had been allowed to conduct into the blue phase, then there would be a sudden jump in output voltage because of the voltage difference between the red and blue phases. Therefore conduction is made to start at points A, B, C, and D, and also A′, B′, C′, and D′; and the variation in conduction will be from the above points to 120° earlier. The appropriate trigger pulses are shown in Fig. 8. The hatched portions of the trigger blocks are the periods during which the thyristors conduct for the motoring condition.

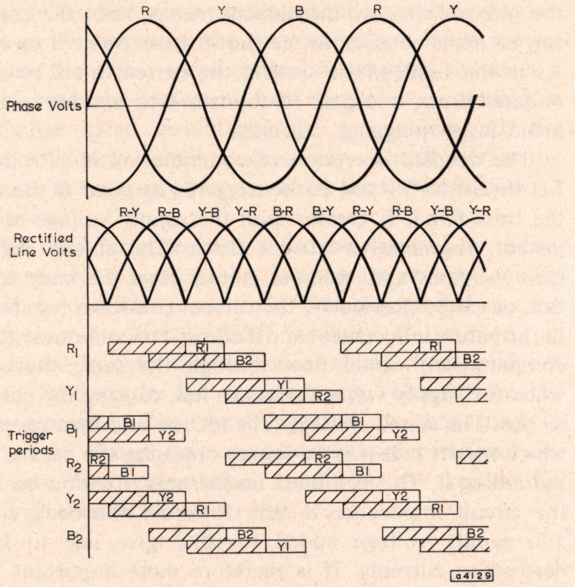


Fig. 8—Trigger requirements in three-phase fully controlled bridge circuit with cross-coupling

For the regeneration condition the circuit and waveforms are shown in Figs. 9 and 10. Following the same reasoning as for the single-phase system (that is, the regeneration current must flow against an opposing supply voltage) we see that the operation of a three-phase regeneration system is as follows.

If the back e.m.f. is as shown at (a), and at point A thyristors R_2 and B_1 are triggered, the back e.m.f. is greater than the voltage between the red and blue phases, therefore current flows back into the mains. The supply voltage is rising, and at point O the back e.m.f. and the supply voltage are equal, but the current continues to flow because of the armature inductance. The current continues until the shaded area above the back e.m.f. line (a) is equal to the shaded area below; that is, the volt-seconds are equal. This particular situation gives discontinuous current.

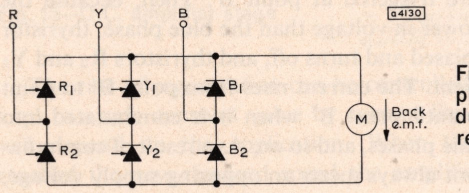


Fig. 9—Three-phase circuit during regeneration

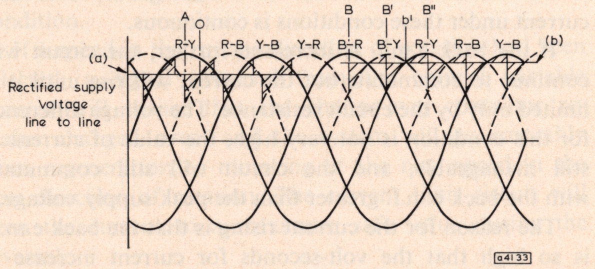


Fig. 10—Three-phase voltage waveforms during regeneration, for two values of back e.m.f.

If the back e.m.f. is increased by, for example, driving the motor faster by mechanical means, then the current can be made continuous, as shown from point B onward. As in the single-phase system, the current must be commutated from one pair of thyristors to another pair to avoid large 'plugging' currents.

The detailed operation of commutation is as follows. Let thyristors Y_1 and B_2 be triggered at point B. Because the back e.m.f. is greater than the supply voltage at this instant, regenerative current flows. The supply voltage then increases and becomes greater than the back e.m.f. But, as stated previously, the current continues because of the armature inductance; and if allowed to continue without commutation, would flow through the same thyristors when the supply voltage starts to fall, causing the current to rise. The supply voltage falls to zero and then reverses, which means that it is no longer opposing the back e.m.f. but aiding it. This is a most undesirable situation because the circuit impedance is very low and the back e.m.f. and supply voltage added together give rise to large destructive currents. It is therefore most important that the current is diverted or commutated to another phase where the supply voltage is opposing.

This is achieved in a similar manner to that used in the single-phase case; that is, by triggering another pair of thyristors at which the supply voltage is lower, causing the current to be diverted. In this particular case, thyristors R_2 and Y_1 are triggered at point B'. Then, because the red phase is lower in voltage than the blue phase, thyristor B_2 is reverse biased and turns off, and thyristors R_2 and Y_1 carry the current. The current rises from point B' to point b', then falls until point B'' when it is commutated into the red and blue phases, and so on. As a result of commutation the current always meets an opposing supply voltage, and its value is maintained at a manageable level. The current under these conditions is continuous.

If the back e.m.f. is increased further, the circuit will continue to commutate but the current will rise until it is limited only by the circuit resistance. The voltage difference for this condition is not very high, the value of current is still manageable, and the circuit will still commutate with the back e.m.f. greater than the peak supply voltage.

The reason for the current rising is that the back e.m.f. is so high that the volt-seconds for current increase is much greater than the volt-seconds for current decrease, therefore the overall current increases. This is shown graphically in Fig. 11. Under condition (a), area A is equal to area B, and the average current is constant. This is a discontinuous current condition. Under condition (b), area A exceeds area B, and the average current is rising. This is a continuous current condition. Under condition (c), area B is reduced to zero, and the current rises to

$$I_{\text{supply}} = \frac{\text{supply voltage}}{\text{circuit resistance}}.$$

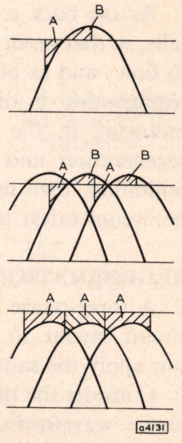


Fig. 11—Three-phase current waveforms

The trigger pulses to cover both motoring and regeneration on three-phase mains are as shown in Fig. 8. It is essential when regenerating under continuous current conditions, that the trigger pulses are not allowed to disappear completely, either one at a time or all together, because this would mean that the current would cease to commutate, with disastrous results. The practice is either to feed permanent pulses to the thyristors at the end of each period (these pulses are called 'end stops') or to ensure that the normal trigger pulses are still just there when fully turned down.

SUMMARY OF REGENERATION ACTION

The reversing action of a d.c. motor system with regeneration is as follows. If the speed control of such a system is suddenly reversed, the trigger pulses are retarded and the motor reversing contactor operates. Then the trigger pulses advance to give the maximum regeneration that current limiting allows; and, as the motor slows down, the trigger pulses advance to keep up the maximum value of regenerative current. As the back e.m.f. falls, the trigger pulses (still controlled by current limitation) advance into the driving range of the bridge and consequently accelerate the motor up to the new selected speed.

An expression for the regenerative current is given in the Appendix (page 290).

Practical Aspects of Regeneration in D.C. Motors

The detailed operation of regeneration in d.c. motors has been discussed in the first section. The present section gives added information in the way of circuitry to enable the principles of d.c. regeneration to be put into practice.

The control circuits for such a system are largely determined by the application, and therefore these will be discussed in general terms only. However, details of a fully toleranced trigger circuit are given together with information on the power circuit.

SYSTEM REQUIREMENTS

The requirements of a d.c. motor control system will depend largely on the application. The following represent, in broad terms, four typical systems in order of complexity.

(i) Simple speed control, one direction of rotation, no electrical braking, open or closed loop.
Applications: fans, pumps, simple industrial machinery, etc.

(ii) Speed control, two directions of rotation, no electrical braking, open or closed loop.
Applications: machine tools, simple hoists, conveyor belts, etc.

(iii) Speed control, two directions of rotation, regenerative braking in one direction.
Applications: crane and hoist, paper mills, etc.

(iv) Speed control, two directions of rotation, regenerative braking in both directions.
Applications: traction, lifts, rolling mills, etc.

The trigger circuit described in a later section can be used for any of these applications. Three-phase thyristor power circuits which can be used with the trigger circuit to satisfy the foregoing requirements are shown in Figs. 12 to 14.

Fig. 12 is the most simple but is not a completely static system owing to the reversing contactor. However, this contactor can be made to operate at zero contact current by ensuring than no thyristor is conducting when change-over takes place. This greatly increases the lifetime of the contactor.

Fig. 13 shows a method of reversing the motor connections to the thyristor bridge using four thyristors instead of a contactor. Each thyristor in the reversing bridge must be capable of passing the full motor current, and because of this the circuit of Fig. 14 is probably more economical.

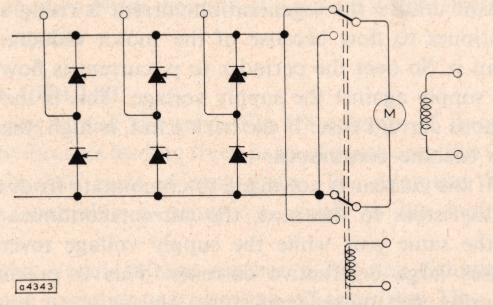


Fig. 12—Simple power circuit with reversing contactor

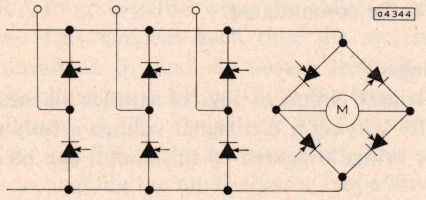


Fig. 13—Power circuit with reversing bridge

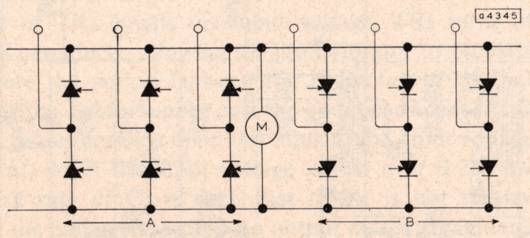


Fig. 14—Power circuit with two bridges

The system of Fig. 14 uses two bridges; bridge A is used for motoring and bridge B for regeneration, for one direction of rotation; and bridge B for motoring and bridge A for regeneration, in the other direction. This means that two trigger circuits are needed, one for each bridge.

Circuit Operation

The operation of the Fig. 12 circuit for application (iv) is as follows. During motoring, the bridge is supplying energy to the motor at a rate determined by the conduction angle of the thyristors and the back e.m.f. of the motor. If, then, the motor speed control is operated to select a lower speed, the motor and load inertia does not allow it to change speed instantaneously, and consequently the speed will decay rather more slowly. To decelerate the motor to the new speed more quickly the system is made to regenerate.

The need for regeneration can be detected by means of the current flowing in the output stage of a d.c. amplifier. If the amplifier is saturated, then the selected speed is lower than the actual speed. Similarly, if the motor is accelerated by external mechanical means the amplifier is turned hard on by the speed-sensing voltage from the tacho-generator. The signal obtained is then used to operate the change-over contactor and set the trigger circuit into the regeneration condition. A typical d.c. amplifier stage is shown in Fig. 15.

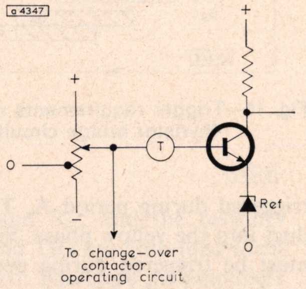


Fig. 15—D.C. amplifier stage

The rate of regeneration can be controlled manually or kept to a safe constant value by current limit. This again will depend on the application. When the motor has reached the new required speed the amplifier comes out of saturation, causing the circuit to revert to the motoring condition.

If a speed in the reverse direction of rotation had been selected, a signal obtained from the control variable resistor will ensure that the change-over contactor remains changed over; and the system will go smoothly from regeneration to motoring at the new speed in the new direction of rotation. The current limit circuit will keep the current to a maximum safe value during the reversal.

One condition can arise which will cause commutation failure and hence large destructive currents. This occurs if the trigger pulses are removed when the circuit is regenerating in the continuous current condition. The condition is explained in the section on *Regeneration Requirements.* It is quite in order to remove the pulses if the circuit is in the discontinuous current state.

POWER CIRCUIT TRIGGER REQUIREMENTS

The trigger requirements of the three-phase fully controlled thyristor bridge for motoring and regeneration operation are shown in Fig. 16. The trigger pulses are cross-coupled within the trigger circuit. This cross-coupling is not necessary for the motoring condition but is essential for regeneration.

Motoring Requirements

For motoring, thyristors R_1 and Y_2 (Fig. 16) must be

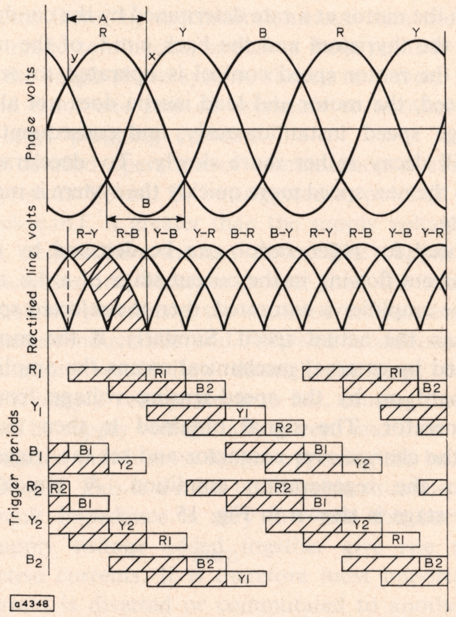


Fig. 16—Trigger requirements of three-phase fully controlled thyristor bridge circuit with cross-coupling

triggered during period A. Then the red phase will conduct into the yellow phase. Similarly, thyristors R_1 and B_2 must be triggered during period B. Then the red phase conducts into the blue phase, and so on.

In period A, zero conduction angle is at point x. To increase the output of the bridge the conduction angle must be increased; and it is maximum when advanced 120° from x to y. From Fig. 16 it can be seen that the motoring trigger requirements can be obtained without cross-coupling. That is, only pulses appropriate to their own phases are required. But, as cross-coupling is required for regeneration, and it does not affect the motoring condition, it can be left in under all conditions.

Regeneration Requirements

The basic requirement of a regeneration system is that energy must be allowed to flow from the load (d.c. motor) back into the mains supply. To effect this transfer of energy the supply voltage must be opposing the flow of current from the motor. This is achieved by triggering the thyristors associated with whichever line voltage is both lower than the back e.m.f. at the point of triggering and is rising in value. This situation is represented in Fig. 17 where thyristors B_1 and R_2 are triggered at point z. From this instant until ω the regeneration current is rising and then continues to flow because of the motor inductance until point p. So over the period z to p, current is flowing into the supply against the supply voltage. This is the discontinuous current case. If the back e.m.f. is high, the current can become continuous.

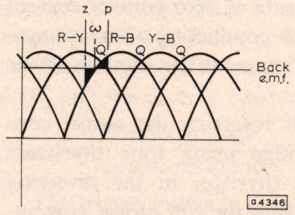

z
p
ω
R–Y
R–B
Y–B
Q
Q
Q
Back e.m.f.
a4346

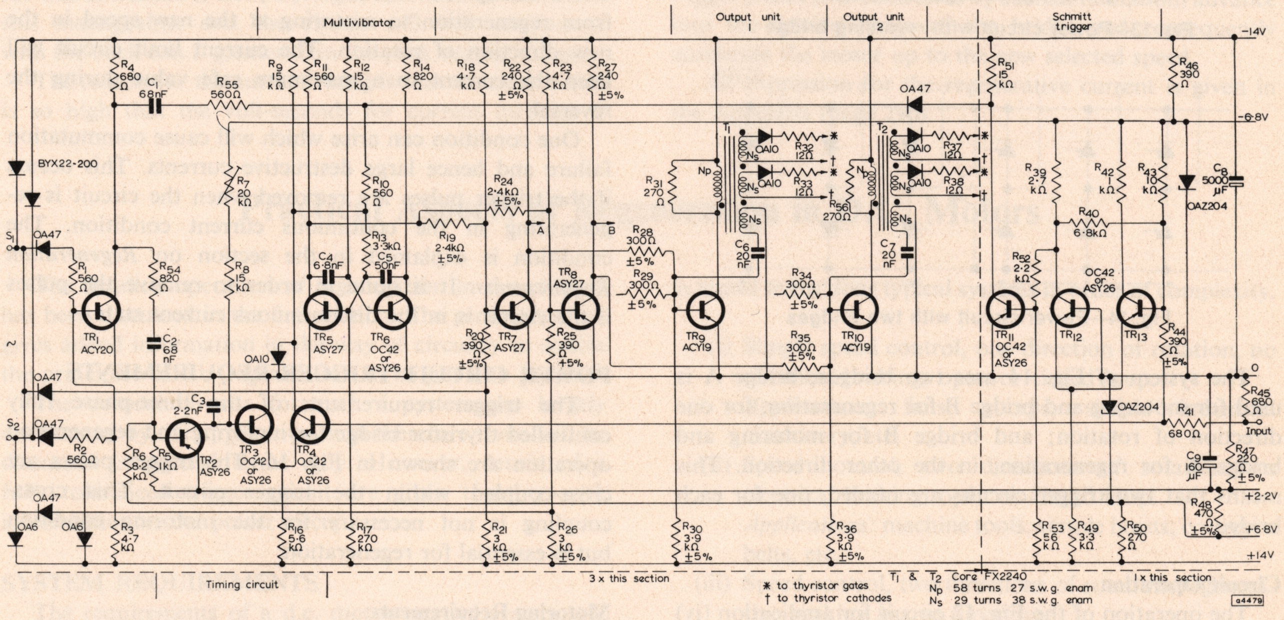
Fig. 17—Commutation

If the current is not made to commutate from one pair of thyristors to the next, the current continues to flow in the same pair while the supply voltage reverses and causes large destructive currents. This is prevented by ensuring that pulses (end stops) always occur just before the crossover points Q (Fig. 17), and this allows the current to be commutated.

Triggering

The trigger circuit of Fig. 18 satisfies all these requirements. By varying a d.c. signal voltage a fully controlled thyristor bridge triggered by this circuit can be controlled from motoring through to regeneration. There is no change-over action in the trigger circuit between the motoring and regeneration conditions, but simply a retardation of the trigger pulses. During the trigger periods, shown hatched in Fig. 16, the circuit is motoring; and during the periods not hatched the circuit is regenerating. The trigger pulses are continuously variable from the end stop pulses through regeneration to maximum output for motoring.

Fig. 19 shows a block diagram of a typical system using the trigger circuit.


Multivibrator
Output unit 1
Output unit 2
Schmitt trigger
Timing circuit
3 x this section
1 x this section
BYX22-200
S1
S2
~
OA47
OA47
OA47
OA47
OA6
OA6
R1 560 Ω
R2 560Ω
R3 4·7 kΩ
R4 680 Ω
C1 68nF
R55 560Ω
R54 820 Ω
C2 68 nF
R6 8·2 kΩ
R5 1kΩ
TR1 ACY20
TR2 ASY26
C3 2·2nF
TR3 OC42 or ASY26
TR4 OC42 or ASY26
R7 20 kΩ
R8 15 kΩ
OA10
R9 15 kΩ
R11 560 Ω
R12 15 kΩ
R14 820 Ω
R10 560 Ω
R13 3·3kΩ
C4 6·8nF
C5 50nF
TR5 ASY27
TR6 OC42 or ASY26
R16 5·6 kΩ
R17 270 kΩ
R18 4·7 kΩ
R22 240 Ω ±5%
R23 4·7 kΩ
R27 240 Ω ±5%
R24 2·4kΩ ±5%
R19 2·4kΩ ±5%
A
B
R20 390 Ω ±5%
TR7 ASY27
TR8 ASY27
R25 390 Ω ±5%
R21 3 kΩ ±5%
R26 3 kΩ ±5%
R28 300Ω ±5%
R29 300Ω ±5%
R30 3·9 kΩ ±5%
R31 270 Ω
Np
Ns
T1
OA10
R32 12Ω
R33 12Ω
C6 20 nF
TR9 ACY19
R34 300Ω ±5%
R35 300Ω ±5%
R36 3·9 kΩ ±5%
T2
R56 270Ω
R37 12Ω
R38 12Ω
C7 20 nF
TR10 ACY19
OA47
R51 15 kΩ
TR11 OC42 or ASY26
R52 2·2 kΩ
R39 2·7 kΩ
R40 6·8kΩ
R42 1 kΩ
R43 10 kΩ
TR12
TR13
OC42 or ASY26
R44 390 Ω ±5%
R46 390 Ω
OAZ204
OAZ204
C8 5000 μF
R41 680Ω
R45 680 Ω
Input
C9 160 μF
R47 220 Ω ±5%
R48 470 Ω ±5%
R53 56 kΩ
R49 3·3 kΩ
R50 270 Ω
−14V
−6·8V
0
+2·2V
+6·8V
+14V
a4479

* to thyristor gates
† to thyristor cathodes

T1 & T2 Core FX2240
Np 58 turns 27 s.w.g. enam
Ns 29 turns 38 s.w.g. enam

Fig. 18—Three-phase phase control circuit with end stops

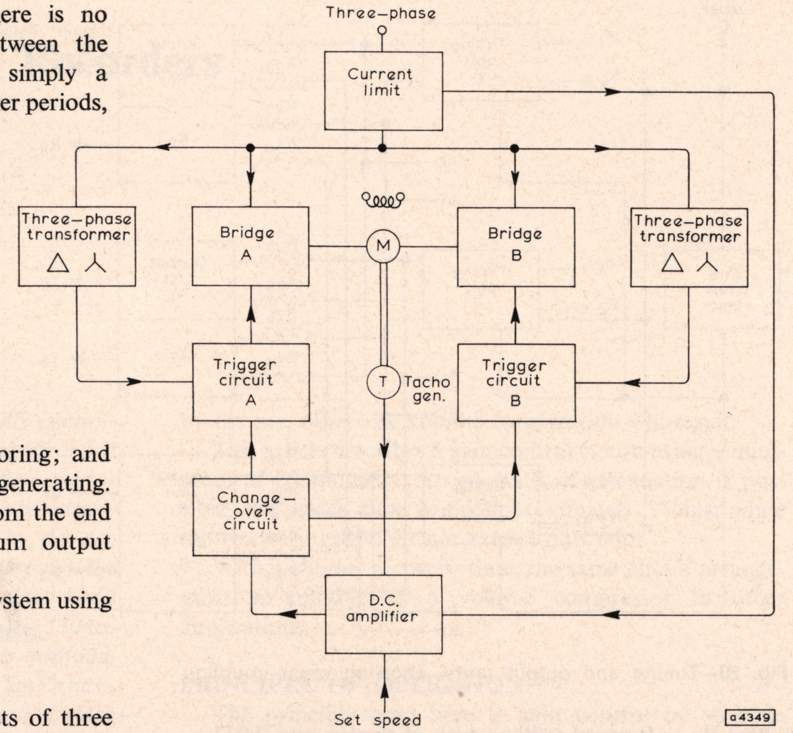

Three–phase
Current limit
Three–phase transformer
Bridge A
M
Bridge B
Three–phase transformer
Trigger circuit A
T
Tacho gen.
Trigger circuit B
Change–over circuit
D.C. amplifier
Set speed
a4349

Fig. 19—Motor control system

TRIGGER CIRCUIT

The trigger circuit shown in Fig. 18 consists of three gated multivibrators feeding three pairs of pulse amplifiers, one for each half cycle of each phase. The operation is as follows, assuming a rising d.c. signal at the input terminal.

When the signal voltage reaches 0·25V the Schmitt trigger circuit switches over, turning off transistor TR_{11}. This removes the clamp from the bistable, TR_5 and TR_6, and allows R_{19} to go negative when the collector of TR_1 goes negative. This happens each time the a.c. supply voltage at terminals S_1 and S_2 passes through zero. Because of the capacitor coupling C_2 to the base of TR_2 from the collector of TR_1, transistor TR_3 of the long-tailed pair TR_3 and TR_4 is made to turn off. This causes TR_4 to turn on, making the multivibrator inoperative.

In this process capacitor C_3 has been charged to $+2{\cdot}2$V and discharges via R_7 and R_8. When the base voltage of TR_3 equals the input voltage, TR_3 turns off and TR_4 conducts, allowing the multivibrator to oscillate. Therefore the period between the beginning of the half cycle of the mains supply and the commencement of the output pulses depends upon the value of the input voltage. However, when the input voltage is just over 0·25V the charging rate of C_3 is such that 0·25V is not reached during the half cycle, and so no output pulses are initiated from this part of the circuit. But when TR_1 cuts off when the mains supply falls to zero (about 8° each side of zero) a pulse is applied to the base of TR_5 of the multivibrator, causing it to emit an output pulse provided that TR_{11} is off, which it is if the input is equal to or greater than 0·25V. This end stop pulse occurs at the end of each half cycle, and since the end stop circuit is common to each phase, all end stop pulses appear and disappear together. This is most important for the correct functioning of commutation in the power circuit under certain conditions, as already explained.

Transistors TR_7 and TR_8 are alternately bottomed for a half cycle by the inputs S_1 and S_2, and the resistors R_{18} and R_{23} ensure that TR_7 and TR_8 are bottomed when the a.c. input to S_1 and S_2 is zero. Therefore the amplifiers receive signals from the multivibrator via TR_7 and TR_8 alternately each half cycle.

The amplifiers are straightforward, with transformer coupled outputs and an additional feedback winding to improve the rise and fall times.

Three timing and output units are required, as indicated at the bottom of Fig. 18. They must be cross-coupled as shown in Fig. 20 to give the trigger periods shown in Fig. 16. This is done by cross-connecting resistors R_{29} and R_{34} of the three units so that when the trigger pulse output amplifier for thyristor R_1 receives input pulses, so does the pulse amplifier for thyristor Y_2. Similarly, R_2 and Y_1, and so on.

The supply voltages for the circuit are obtained from the a.c. synchronising signal at S_1 and S_2. The two BYX22–200 diodes in each circuit supply the −14V supply rail, which is smoothed by the 5000μF capacitor C_8. The two OA6 diodes supply the +14V rail, and the other voltage rails are obtained by potential divider and voltage regulator diodes. As the three sections have their supplies commoned, the overall power supply is as shown in Fig. 21 and is equivalent to half-wave rectification from a six-phase bridge.

The interconnections between the three timing and output units, and the transformer for the a.c. inputs, are given in Figs. 20, 22, and 23. The transformer has to be a delta primary and a star secondary to give the required 30° phase shift to the trigger pulses. The output voltage must be such that when rectified it supplies a mean level of d.c. of 14V unsmoothed for the +14V rail. Therefore:

$$V_{sec} = 0{\cdot}707\left(\frac{14}{0{\cdot}933} + V_F\right)$$

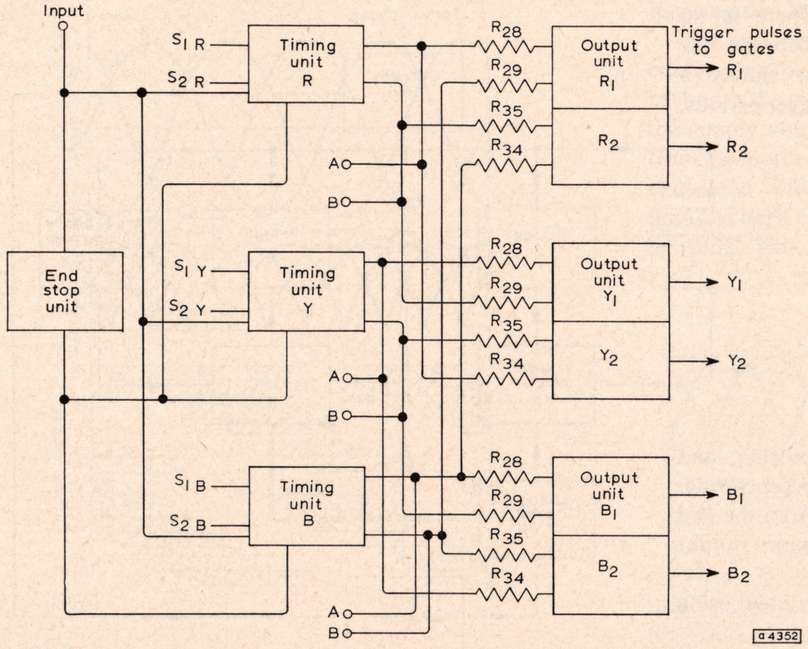

Input
S1 R
S2 R
Timing unit R
R28
R29
R35
R34
Output unit R1
R2
Trigger pulses to gates
R1
R2
A
B
End stop unit
S1 Y
S2 Y
Timing unit Y
Output unit Y1
Y2
Y1
Y2
S1 B
S2 B
Timing unit B
Output unit B1
B2
B1
B2
g4352

Fig. 20—Timing and output units, showing cross-coupling

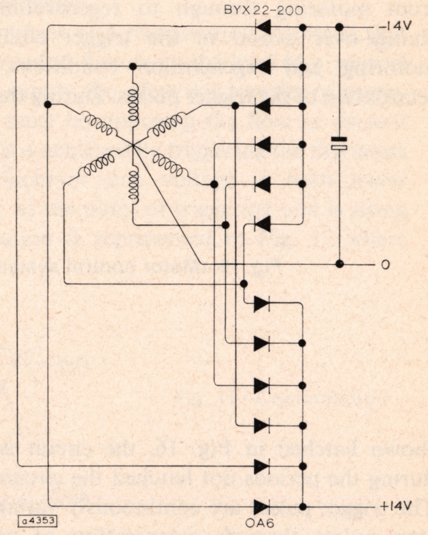

BYX22-200
−14V
0
+14V
OA6
g4353

Fig. 21—Power supply for three-phase phase control circuit

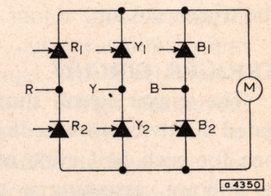

R1
Y1
B1
R
Y
B
M
R2
Y2
B2
g4350

Fig. 22—Three-phase power circuit

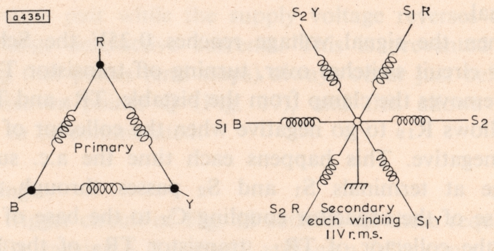

g4351
R
Primary
B
Y
S2 Y
S1 R
S1 B
S2 B
S2 R
S1 Y
Secondary each winding 11V r.m.s.

Fig. 23—Transformer for a.c. inputs

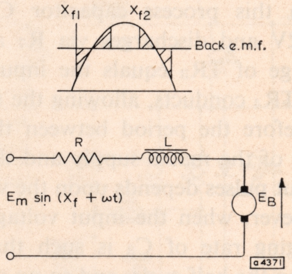

Xf1
Xf2
Back e.m.f.
R
L
Em sin (Xf + ωt)
EB
g4371

Fig. 24—Equivalent circuit

where V_F = forward voltage drop of diodes, and 0·933 = mean level factor of v_{pk} for six-phase half-wave rectifier system. Therefore if $V_F = 0{\cdot}6$, then $V_{sec} = 11V$.

Circuit Characteristics

The circuit is toleranced to a temperature range of 55°C to —20°C with a variation of trigger angle between phases of 5° or less.

The input characteristic is such that nothing happens up to 0·25V. When this value is reached the end stop pulses appear. As the input voltage is increased still further, the conduction angle is increased until maximum conduction angle is reached at approximately 2V. The input impedance is 680Ω.

APPENDIX

Calculation of Instantaneous and Average Currents

An expression for the regeneration current in terms of the circuit constants, the supply voltage, and the back e.m.f. has been derived. It can be used to obtain the motoring current as well as the regeneration current, depending on the values of the trigger angle and the back e.m.f. In Fig. 24, E_m is the maximum value of the supply voltage, X_{f1} is the regeneration trigger angle, and X_{f2} the motoring trigger angle.

$$I = \frac{E_m}{R}\left[-\frac{E_B}{E_m} + \cos\theta \sin(X_f + \omega t - \theta) + \right.$$
$$\left. + \left(\frac{E_B}{E_m} - \cos\theta\sin(X_f - \theta)\exp(-Rt/L)\right)\right] \quad \ldots(1)$$

This expression is identical to that obtained by Vedder and Puchlowski for systems employing thyratrons (Ref. 1).

From definitions,

$$I = 0 \text{ when } \omega t = 0 \text{ and } (X_f + \omega t) = X_s$$

therefore for discontinuous current:

$$\frac{E_B}{E_m} - \cos\theta\sin(X_s - \theta)$$
$$= \left[\frac{E_B}{E_m} - \cos\theta\sin(X_f - \theta)\right]\exp(-Rt/L) \quad \ldots(2)$$

From this, values of X_s can be calculated for various values of X_f; then if these are reinserted in the original equation, values for I can be calculated and hence I_{rms} and I_{av}. It will be appreciated that this is a lengthy calculation and is therefore beyond the scope of this article, but Vedder and Puchlowski produced an equation for I_{av} from the above, as follows:

$$I_{av} = \frac{3E_m}{2\pi R}\left[\cos X_f - \cos X_s - \frac{E_B}{E_m}(X_s - X_f)\right]. \quad \ldots(3)$$

This is again for the discontinuous case. For the continuous case the d.c. component must be taken into consideration.

REFERENCE

1. Vedder, E. H. and Puchlowski, K. P., 'Theory of Rectifier D.C. Motor Drive'. **Trans. A.I.E.E. ('Electrical Engineering'),** Vol. 62, Dec. Supp., 1943, pp. 863-870.

A.G.C. Circuit for Tape Recorders

R. Osborne
Mullard Central Application Laboratory

This article gives a simple circuit for automatically controlling the gain of tape recording amplifiers, such that distortion due to tape saturation may be avoided. The principle used is that of variable attenuation by the variable slope resistance of diodes, the slope resistance being controlled by the d.c. current flowing through them, which is derived from the rectified signal to be controlled. Attenuation of about 40dB may be obtained, the attack time being approximately 150ms. The rate of increase of gain at no time exceeds 6dB/min, and providing the controlled signal is less than 2mVr.m.s., the attenuating network introduces harmonic distortion which does not exceed 2%.

INTRODUCTION

When a tape recorder is recording signals, it is important that the recording current flowing through the recording head does not exceed a certain level, otherwise severe distortion of the recorded signal will result due to tape saturation. One method of ensuring that this does not occur is by monitoring the recording current level by magic eye or meter, and adjusting the recording level manually. Another method is to incorporate a circuit which electronically reduces the gain of the recording amplifier when the optimum recording level is exceeded. So that distortion is recorded for only a very short time, the gain control must respond very rapidly to the overload signal. Once the gain is reduced, the recovery time must be long so that the rate of increase of gain is not detected by the ear, otherwise volume compression will result.

This article describes a gain control circuit using a minimum of components. About 40dB of attenuation is possible, the attack time being approximately 150ms, while the recovery rate does not exceed 6dB/min.

With a shorter recovery time, the same circuit arrangement is suitable as a volume compressor in other applications.

PRINCIPLE OF OPERATION

The principle used here is gain control by variable attenuation. This is produced by varying the slope resistance of a diode by controlling the d.c. current through it. Unfortunately, the exponential nature of the characteristic introduces severe harmonic distortion, thus it is common practice to use two diodes in anti-parallel to the a.c. signals, to reduce this distortion. The present circuit makes use of the base-emitter junction of a transistor (the transistor being required incidentally as a d.c. current amplifier) as one of the diodes, thus saving one diode.

The d.c. controlling signal is derived from the signal level by a rectifier circuit having the required attack and recovery times, and feeding into the transistor which forms part of the variable attenuator. This system is incorporated in the record function amplifier.

CIRCUIT DESCRIPTION

The gain control circuit is given in Fig. 1. Unit A

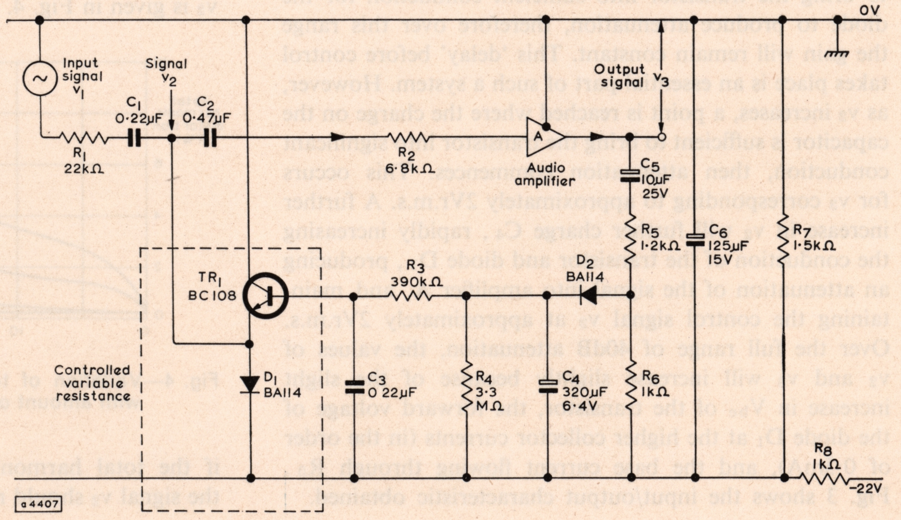

0V
Input signal v1
Signal v2
C1 0·22μF
C2 0·47μF
R1 22kΩ
R2 6·8kΩ
A
Audio amplifier
Output signal v3
C5 10μF 25V
R5 1·2kΩ
C6 125μF 15V
R7 1·5kΩ
TR1 BC108
R3 390kΩ
D2 BA114
Controlled variable resistance
D1 BA114
C3 0 22μF
R4 3·3 MΩ
C4 320μF 6·4V
R6 1kΩ
R8 1kΩ
−22V
a4407

Fig. 1—Gain control

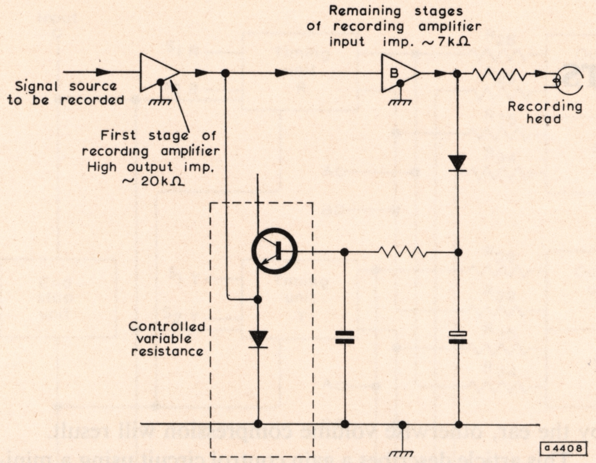


Fig. 2—Incorporation of gain control circuit in tape amplifier

is an amplifier which may be part of the recording amplifier (unit B in Fig. 2). The input impedance R_2 of amplifier A should be about 7kΩ. The input signal v_1 passes to the amplifier via a high resistance (about 20kΩ). Gain control is achieved by the shunting effect of the slope resistance of the diode D_1 in a.c. anti-parallel with the slope resistance of the base-emitter junction of the transistor TR_1 as described above. Capacitor C_3 satisfies the condition that the diodes are in anti-parallel for a.c. signals.

The long recovery time is provided by C_4 being discharged through the transistor by R_3. Over a wide range of the attenuation, R_3 is large compared with the input impedance of the transistor plus diode, and the rate of change of gain in dB/min is constant. However, at very low transistor collector currents the input impedance becomes significant. Resistor R_4 is included to reduce this effect.

The attack time is determined primarily by C_4 and the source impedance of the charging circuit, in this case R_5 in parallel with R_6. For attack times of about 200ms, this source impedance should be approximately 600Ω.

The circuit operates as follows. For very small signal input levels (v_2), the rectification of the control signal v_3 charges up C_4 to the peak value of v_3 attenuated by R_5 and R_6. However, this charge on the capacitor is insufficient to bring the transistor into sufficient conduction for the diode to produce attenuation, therefore over this range the gain will remain constant. This 'delay' before control takes place is an essential part of such a system. However, as v_2 increases, a point is reached where the charge on the capacitor is sufficient to bring the transistor into significant conduction, then attenuation commences. This occurs for v_3 corresponding to approximately 2Vr.m.s. A further increase in v_2 will further charge C_4, rapidly increasing the conduction of the transistor and diode D_1, producing an attenuation of the signal into amplifier A, and maintaining the control signal v_3 at approximately 2Vr.m.s. Over the full range of 40dB attenuation, the values of v_2 and v_3 will increase slightly because of the slight increase in V_{be} of the transistor, the forward voltage of the diode D_1 at the higher collector currents (in the order of 0·3mA), and the base current flowing through R_3. Fig. 3 shows the input/output characteristic obtained.

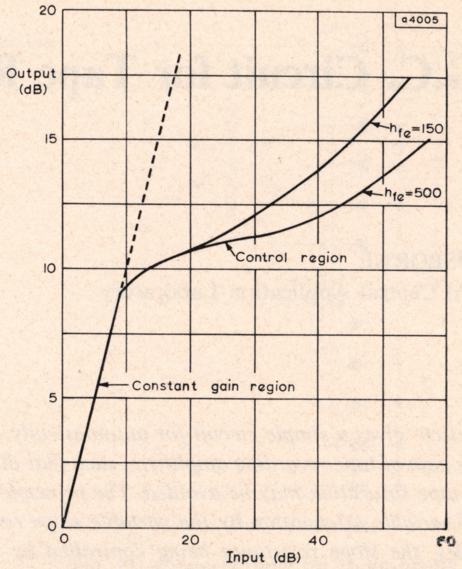


Fig. 3—Variation of output with input before, and in the control region, for high and low gain transistors

The transistor used is the BC108 epitaxial silicon planar n.p.n. transistor, which with its low leakage current and high gain, makes this simple arrangement possible. Silicon diodes are used, also for their low leakage current properties.

Fig. 2 indicates how the circuit is likely to be incorporated into a recording system. If the attenuation were performed at the input of the system, the signal level would be controlled, but the noise originating in the first transistor would not be reduced. It would be preferable to have the attenuation after the first stage as indicated, reducing the noise and signal simultaneously. A high voltage level is required to operate the attenuating circuit, thus it is probable that the only stage where this is available is that stage which provides the recording current to the head.

PERFORMANCE

Harmonic Distortion

The variation of total harmonic distortion of the signal v_2 with the amount of attenuation for various values of v_2 is given in Fig. 4. It can be seen from these curves that if the total harmonic distortion is not to exceed 2%, the signal v_2 should not exceed 4mVr.m.s.

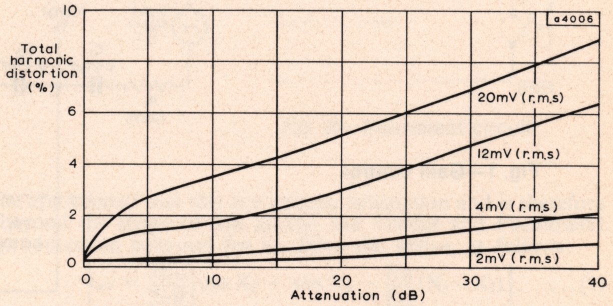


Fig. 4—Variation of total harmonic distortion of signal V_2 with amount of attenuation for various values of V_2

Control Region

As explained earlier, in the region where the circuit controls the gain, the signal level is not maintained exactly constant and independent of the amount of attenuation. In Fig. 3, the increase of output level with attenuation is shown for both high gain and low gain transistors. In the case of the low gain transistors, it can be seen that for 40dB attenuation the output level has increased 7dB whereas the corresponding figures for the high gain case are 4·5dB increase in output signal for 40dB attenuation.

Recovery Time

The reduction of attenuation with time for high and low gain transistors is given in Fig. 5. It can be seen that the rate of change of attenuation with time is substantially constant over a large portion of the range, and at no time exceeds a rate of 6dB/min. Also the rate is almost independent of the gain of the transistor.

USE OF ELECTROLYTIC CAPACITORS IN TIMING CIRCUITS

The maximum leakage currents quoted in published data on electrolytics are too high for use in long time-constant circuits. However, experience on capacitors taken off the shelf after varying storage periods suggests that, in practice, the leakage current is adequately low, or falls to adequately low values after a relatively short period of use.

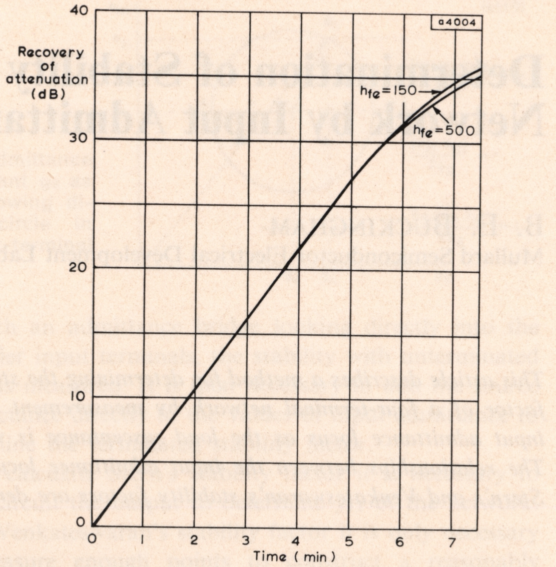


Fig. 5—Decrease of attenuation with time for high and low gain transistors

This leads us to conclude that electrolytic capacitors of suitable type are readily available. At the worst, a tape recorder which has been stored for a very long time, will have an initial recovery time slightly shorter than normal, but this will rapidly approach the normal value.

Determination of Stability Factor of a Four-terminal Network by Input Admittance Measurements

B. E. Buckingham
Mullard Semiconductor Electrical Development Laboratory

This article describes a method for determining the stability factor of a four-terminal network by measurement of the input admittance locus as the load susceptance is varied. The relationships between the input admittance locus and Stern's and Venkateswaran's stability factors are derived.

INTRODUCTION

A method of determining the stability factor of an amplifier from input admittance measurements is described. The advantage of this method is that the stability factor of the whole amplifier stage including strays is determined, and this may be very different from that calculated from transistor y-parameters. This is particularly important at u.h.f. where the strays are large and have a great significance on transistor performance.

DEFINITIONS OF STABILITY FACTOR

Until recently Stern's stability factor was perhaps the most frequently used measure of stability for an amplifier. This is defined, for an unterminated four terminal network, as:

$$S = 2g_{11}g_{22}/[R_e(y_{12}y_{21})+|y_{21}y_{12}|]. \qquad \ldots(1)$$

Or, when the network is terminated, that is when embedded in the amplifier circuit, the working stability is:

$$S_s = 2g_1g_2/[R_e(y_{12}y_{21})+|y_{21}y_{12}|] \qquad \ldots(2)$$

where

$$g_1 = g_{11}+g_s \text{ and } g_2 = g_{22}+g_L$$

g_s and g_L being the source and load conductances respectively.

Stern's stability factor is a measure of nearness to oscillation and suffers from the disadvantage that although the circuit may be well away from oscillation, and have a high stability factor, there may still be a large amount of interaction between input and output.

An alternative stability factor which does give a measure of the interaction between input and output circuits is that proposed by Venkateswaran (Ref. 1):

$$S_v = \eta+(\eta^2-1)^{\frac{1}{2}} \qquad \ldots(3)$$

where

$$\eta = [2g_{11}g_{22}-R_e(y_{21}y_{12})]/|y_{21}y_{12}| \qquad \ldots(4)$$

and where η is in itself a measure of stability. Since η can be more easily compared with S_s in terms of the y-parameters, this will be considered rather than S_v.

For a terminated network the working stability factor is:

$$\eta_t = [2g_1g_2-R_e(y_{21}y_{12})]/|y_{21}y_{12}|. \qquad \ldots(5)$$

CALCULATION OF INPUT ADMITTANCE LOCUS, WITH VARIATION OF LOAD SUSCEPTANCE

The input admittance of a four-terminal network is given by:

$$y_{in} = y_{11}-y_{21}y_{12}/(y_{22}+y_L) \qquad \ldots(6)$$

or

$$y_{in} = g_{11}+jb_{11}-(R+jI)/(g_2+jb_2) \qquad \ldots(7)$$

where

$$R = R_e(y_{21}y_{12}) \quad g_2 = g_{22}+g_L$$
$$I = I_m(y_{21}y_{12}) \quad b_2 = b_{22}+b_L.$$

Rationalising Eq. 7:

$$g_{in}+jb_{in} = g_{11}+jb_{11}-[R.g_2+I.b_2+j(I.g_2-R.b_2)]/ /(g_2{}^2+b_2{}^2). \qquad \ldots(8)$$

Let $x = g_{in}-g_{11}$ and $y = b_{in}-b_{11}$.

Then from Eq. 8:

$$x = -(R.g_2+I.b_2)/(g_2{}^2+b_2{}^2) \qquad \ldots(9)$$
$$y = -(I.g_2-R.b_2)/(g_2{}^2+b_2{}^2). \qquad \ldots(10)$$

Eliminating b_2 from Eqs. 9 and 10:

$$x/y = (R.g_2+I.b_2)/(I.g_2-R.b_2)$$
$$I.g_2.x-R.b_2.x = R.g_2.y+I.b_2.y$$
$$b_2 = g_2(I.x-R.y)/(I.y+R.x). \qquad \ldots(11)$$

Substituting for b_2 in Eq. 9:

$$x\{g_2{}^2+g_2{}^2[(I.x-R.y)/(I.y+R.x)]^2\} = = -\{R.g_2+I.g_2[(I.x-R.y)/(I.y+R.x)]\}$$

which simplifies to

$$g_2(x^2+y^2)+R.x+I.y = 0. \qquad \ldots(12)$$

Finally, resubstituting for x and y in Eq. 12:

$$g_2(g_{in}-g_{11})^2+g_2(b_{in}-b_{11})^2+R.(g_{in}-g_{11})+ +I(b_{in}-b_{11}) = 0. \qquad \ldots(13)$$

This is the equation of a circle in the g_{in}, b_{in} plane with centre at

$$g_{in} = g_{11} - \frac{R}{2g_2} \text{ and } b_{in} = b_{11} - \frac{I}{2g_2}.$$

The diameter of the circle is

$$[(R/g_2)^2+(I/g_2)^2]^{\frac{1}{2}} \text{ or } |y_{21}y_{12}|/g_2.$$

A family of circles exists, each passing through the point g_{11}, b_{11} and whose diameter depends on the value of g_2. In the case where b_2 is constant and g_2 is allowed to vary, a family of circles is obtained which is orthogonal to the family of circles for constant g_2. That is with centre

$$g_{in} = g_{11} - \frac{I}{2b_2} \text{ and } b_{in} = b_{11} - \frac{R}{2b_2}$$

and the diameter $|y_{21}y_{12}|/b_2$.

Since b_2 can be either positive or negative, two families of circles exist, which are cotangential at the point g_{11}, b_{11} as shown in Fig. 1.

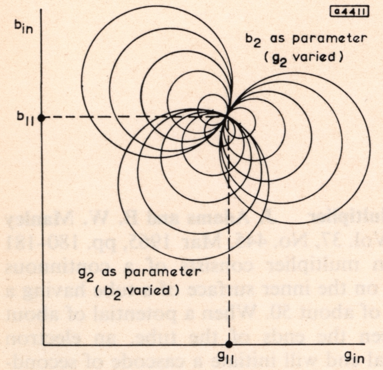


Fig. 1—Locus of input admittance of four-terminal network

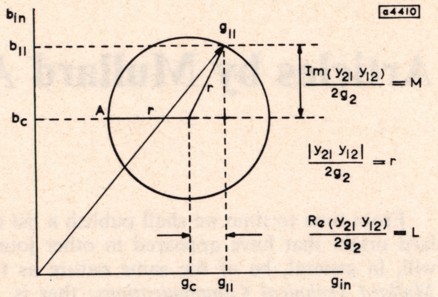


Fig. 2—Input admittance circle for constant g_2 as b_2 is varied, showing dimensions of circle in terms of four-terminal parameters

INPUT ADMITTANCE LOCUS AND ITS RELATIONSHIP TO STABILITY FACTOR

A typical input admittance circle for constant g_2 is shown in Fig. 2.

Putting Stern's stability factor in terms of the dimensions of Fig. 2:

$$\begin{aligned} S_s &= 2g_1g_2/[R_e(y_{21}y_{12})+|y_{21}y_{12}|] \\ &= g_1/\left(R_e\frac{(y_{21}y_{12})}{2g_2}+\frac{|y_{21}y_{12}|}{2g_2}\right) \\ &= g_1/(L+r). \qquad \ldots(14) \end{aligned}$$

Putting Venkateswaran's stability factor in terms of the dimensions of Fig. 2:

$$\begin{aligned} \eta &= [2g_1g_2-R_e(y_{21}y_{12})]/|y_{21}y_{12}| \\ &= [g_1-R_e(y_{21}y_{12})/2g_2]/[|y_{21}y_{12}|/2g_2] \\ &= (g_1-L)/r \end{aligned}$$

or

$$\begin{aligned} \eta &= (g_s+g_{11}-L)/r \\ &= (g_s+g_c)/r. \qquad \ldots(15) \end{aligned}$$

From Eqs. 14 and 15 the difference between the stability factors can readily be seen. Stern's stability factor is dependent on the position of the vector y_{11} relative to the circle; and should y_{11} be at A for the circle shown, then S_s would be infinite.

Venkateswaran's stability factor, however, is independent of the position of the vector y_{11} on the circle and depends only on the radius of the circle compared with the distance of the centre of the circle from the b- axis.

It should be noted that adding g_s to g_c or g_{11} shifts the circle horizontally to the right by a distance g_s and therefore increases the stability.

The sign of $I_m(y_{21}y_{12})$ has no effect on the stability factor, since it makes no difference if the centre of the circle is above or below b_{11}. However, the sign of $R_e(y_{21}y_{12})$ is important. If it is positive, the centre of the circle lies to the left of g_{11}, and if it is negative, the centre lies to the right of g_{11}. Therefore the sign of L is accounted for in Stern's equation by taking the distance between g_{11} and the minimum value of g_{in}.

MEASUREMENT OF STABILITY FACTOR OF R.F. AMPLIFIER

From the foregoing it may be seen that measurement of the change in input admittance with load susceptance provides a simple method for the determination of amplifier stability.

With an admittance bridge looking directly into the amplifier input terminals, the stability with unterminated input is measured. It is then necessary to add the source conductance directly to the values of input admittance to determine the terminated stability factor.

To calculate Stern's stability factor it is necessary to short-circuit the output to determine y_{11} of the network. With Venkateswaran's stability factor it is only necessary to measure enough points to construct a reasonably accurate circle, and then to read off the lowest and highest values of conductance.

If these values are g_a and g_b, then the stability factor with unterminated input is the mean value divided by half the difference. Therefore

$$\eta = (g_b+g_a)/(g_b-g_a)$$

and the terminated stability factor

$$\begin{aligned} \eta_t &= \left(\frac{g_b+g_a}{2}+g_s\right)\bigg/\left(\frac{g_b-g_a}{2}\right) \\ &= (g_b+g_a+2g_s)/(g_b-g_a). \end{aligned}$$

SYMBOLS

S_s = Stern's stability factor
S_v = Venkateswaran's stability factor (published)
η = Venkateswaran's stability factor (modified)
$y_{11} = g_{11}+jb_{11}$ = input admittance of four-terminal network with output short-circuit
$y_{12} = g_{12}+jb_{12}$ = reverse transfer admittance of four-terminal network with input short-circuit
$y_{21} = g_{21}+jb_{21}$ = forward transfer admittance of four-terminal network with output short-circuit
$y_{22} = g_{22}+jb_{22}$ = output admittance of four-terminal network with input short-circuit
$R = R_e(y_{21}y_{12})$ = real part of transfer admittance product
$I = I_m(y_{21}y_{12})$ = imaginary part of transfer admittance product
$g_1 = g_{11}+g_s$ = total input damping
$g_2 = g_{22}+g_L$ = total output damping
$b_1 = b_{11}+b_s$ = total input susceptance
$b_2 = b_{22}+b_L$ = total output susceptance
$y_{in} = g_{in}+jb_{in}$ = input admittance of network with given load
g_s = source conductance
g_L = load conductance
r = radius of stability circle

ACKNOWLEDGEMENT

Acknowledgement is made to J. C. Beckley of the Mullard Central Application Laboratory, who first introduced this technique as a method of comparing amplifier stability.

REFERENCE

Venkateswaran, S. 'An Invariant Stability Factor and its Physical Significance'. **Proc. I.E.E.**, Part C, Vol. 109, No. 15, March 1962, pp. 98 to 102.

Articles by Mullard Authors

From time to time we shall publish a list of articles of Mullard origin that have appeared in other journals. The articles will, in general, be of the same nature as those published in *Mullard Technical Communications;* that is, they will be concerned with applications of valves, cathode ray tubes, semiconductor devices, and other active or passive components used in 'entertainment' or 'industrial' electronics. Articles on fundamental research, device development, manufacturing technology, education, economics, and so on, will not normally be included.

The Metal Oxide Semiconductor Transistor
R. A. Hilbourne and J. F. Miles
Electronic Engineering, Vol. 37, No. 445, Mar 1965, pp. 156–160

The metal oxide semiconductor transistor is a new semiconductor device working on the field effect principle, but offering a much higher, temperature stable, input resistance. The development has now reached the point where devices with stable characteristics are commercially available, and this article is intended to serve as an introduction to the device.

The construction of the m.o.s.t. is described, and the theory of its operation, and its characteristics are given. This is followed by a description of several applications of the device both in linear circuits and as a switch.

Using the Cold Cathode Tube **M. A. MacDougall**
Electronics, Vol. 38, No. 6, 22 Mar 1965, pp. 78–82

Improved glow tubes are finding new applications in display, and are performing many duties normally associated with more sophisticated devices.

Cold Cathode Tubes as Triggers **M. E. Bond**
Electronics, Vol. 38, No. 7, 5 Apr 1965, pp. 76–85

Subminiature tubes are finding new uses in low speed logic and relay circuits. Inexpensive, but highly reliable, they have a wide technical application.

Using Cold Cathode Tubes to Count and Store **G. F. Jeynes**
Electronics, Vol. 38, No. 8, 19 Apr 1965, pp. 80–89

New low voltage tubes can be used in logic systems in industrial controls, instrumentation, or on the production line.

Picture Tubes for Television Displays using Quadruple Scan Magnification **K. E. Johnson**
The Radio and Electronic Engineer, Vol. 28, No. 2, Aug 1964, pp. 115–128

The design and performance of picture tubes having 8½ in screens, 90 deg. deflection angle, and using permanent magnet quadruple scan magnifying lenses is described. These tubes require only a thirtieth or less of the line scanning power of a conventional tube having the same final anode voltage and deflection angle. Field scanning power is also slightly reduced.

Description of both magnetic and electrostatic focusing systems are given. The magnetically focused tube suffers from the disadvantages that it requires an extremely stable e.h.t. supply, and a large number of focusing adjustments. The electrostatically focused tube requires no more stabilisation of the e.h.t. than is usually required for conventional tubes and reduces the number of focusing adjustments considerably.

However, the increased length of the tube (40% greater than conventional tubes) and increased complexity of manufacture and adjustment compared with conventional tubes coupled with the need for a separate e.h.t. and video h.t. supply generator, has led the author to the conclusion that this technique has little to offer at the present stage of picture tube development.

The Channel Electron Multiplier **J. Adams and B. W. Manley**
Electronic Engineering, Vol. 37, No. 445, Mar 1965, pp. 180–181

The channel electron multiplier consists of a continuous resistive dynode formed on the inner surface of a tube having a ratio length to diameter of about 50. When a potential of about 2·5kV is applied between the ends of the tube, an electron entering the low potential end will initiate a cascode of secondary electrons from the wall. The number of electrons emerging from the high potential end of the tube may exceed 10^5.

Such devices may be small and rugged, and are potentially useful in space applications. Various channel geometries have been investigated at the Mullard Research Laboratories, and it has been found that by making the channel multiplier of spiral form, pressure effects are avoided completely.

Trigger Mechanisms and Sensitivity in Cold Cathode Tubes
G. F. Weston
Proc. I.E.E., Vol. 111, No. 2, Sep 1964, pp. 1546–1550

Although cold cathode trigger tubes are widely used in computer and control circuits, their firing mechanism is not completely understood, particularly when a capacitor is connected between the trigger and the cathode. The paper describes an investigation of the firing mechanism with d.c. input circuits of high impedance.

The effect of the trigger current on the anode breakdown has also been investigated. At onset of oscillation, there is a large reduction in the anode breakdown voltage. Thereafter increasing the mean trigger current has only a small effect. This is particularly true of large capacitances. As a result it is possible to describe the sensitivity in terms of the trigger capacitance needed to lower the anode breakdown voltage to a given value.

An Appraisal of some Decoding Systems for Three Gun N.T.S.C. Colour Television Receivers **D. R. Birt and K. G. Freeman**
The Radio and Electronic Engineer, Vol. 28, No. 1, July 1964, pp. 33–53

Results are given of some practical and theoretical investigations of several types of decoding systems for use with three gun N.T.S.C. receivers. It is concluded that the (R−Y)/(B−Y) type of decoder is the simplest and most convenient from the point of view of design and adjustment, and is likely to be the least critical in operation.

Transistors in Television **D. G. Thompson**
(1) to (5), see list published in Vol. 8, No. 71, Aug 1964, pp. 31–32

(6) Video Amplifiers and AGC systems
Electrical and Radio Trading, Vol. 63, No. 804, 6 Aug 1964, pp. 33–34
(7) Non linear Switching Circuits
ibid, Vol. 63, No. 811, 24 Sep 1964, pp. 68, 77
(8) Line Output Stage Circuit Requirements
ibid, Vol. 64, No. 816, 29 Oct 1964, pp. 35–37
(9) A Closer Look at Line Output and Boost H.T. Circuits
ibid, Vol. 64, No. 819, 19 Nov 1964, pp. 42–43
(10) Line Oscillators
ibid, Vol. 64, No. 822, 10 Dec 1964, pp. 47–49
(11) Phase Detectors and Flywheel Sync.
ibid, Vol. 64, No. 823, 17 Dec 1964, pp. 21–22
(12) Field Timebases
ibid, Vol. 65, No. 828, 28 Jan 1965, pp. 29–30
(13) Dual Standard Circuits
ibid, Vol. 65, No. 834, 11 Mar 1965, pp. 39–42
(14) Prospects and Power Circuits (Final article)
ibid, Vol. 65, No. 837, 1 Apr 1965, pp. 28, 33